

HIGH-SPEED LINK MODELING:  
ANALOG/DIGITAL EQUALIZATION AND MODULATION TECHNIQUES

A Thesis

by

KEYTAEK LEE

Submitted to the Office of Graduate Studies of  
Texas A&M University  
in partial fulfillment of the requirements for the degree of

MASTER OF SCIENCE

May 2012

Major Subject: Electrical Engineering

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## ABSTRACT

High-speed Link Modeling: Analog/Digital Equalization and Modulation Techniques.

(May 2012)

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High-speed serial input-output (I/O) link has required advanced equalization and modulation techniques to mitigate inter-symbol interference (ISI) caused by multi-Gb/s signaling over band-limited channels. Increasing demands for transceiver power and area complexity has leveraged on-going interest in analog-to-digital converter (ADC) based link, which allows for robust equalization and flexible adaptation to advanced signaling. With diverse options in ISI control techniques, link performance analysis for complicated transceiver architectures is very important. This work presents advanced statistical modeling for ADC-based link, performance comparison of existing modulation and equalization techniques, and proposed hybrid ADC-based receiver that achieves further power saving in digital equalization.

Statistical analysis precisely estimates high-speed link margins at given implementation constraints and low target bit-error-rate (BER), typically ranges from  $10^{-12}$  to  $10^{-15}$ , by applying proper statistical bound of noise and distortion. The proposed statistical ADC-based link modeling utilizes bounded probability density function (PDF) of limited quantization distortion (4-6 bits) through digital feed-forward and decision

feedback equalizers (FFE-DFE) to improve low target BER estimation. Based on statistical modeling, this work surveys the impact of insufficient equalization, jitter and crosstalk on modulation selection among two and four level pulse amplitude modulation (PAM-2 and PAM-4, respectively) and duobinary, and ADC resolution reduction performance by partial analog equalizer (PAE).

While the information of channel loss at effective Nyquist frequency and signaling constellation loss initially guides modulation selection, the statistical analysis results show that PAM-4 best tolerates jitter and crosstalk, and duobinary requires the least equalization complexity. Meanwhile, despite robust digital equalization, high-speed ADC complexity and power consumption is still a critical bottleneck, so that PAE is necessitated to reduce ADC resolution requirement. Statistical analysis presents up to 8-bit resolution is required in 12.5Gb/s data communications at 46dB of channel loss without PAE, while 5-bit ADC is enough with 3-tap FFE PAE. For optimal ADC resolution reduction by PAE, digital equalizer complexity also increases to provide enough margin tolerating significant quantization distortion. The proposed hybrid receiver defines unreliable signal thresholds by statistical analysis and selectively takes additional digital equalization to save potentially increasing dynamic power consumption in digital. Simulation results report that the hybrid receiver saves at least 64% of digital equalization power with 3-tap FFE PAE in 12.5Gb/s data rate and up to 46dB loss channels. Finally, this work shows the use of embedded-DFE ADC in the hybrid receiver is limited by error propagation.

## DEDICATION

*To my wife and two sons,  
for giving me the reason and hope of my life*

*To my parents and parents in laws,  
for their love and support*

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## CHAPTER I

### INTRODUCTION

As the data rate of high-speed input/output (I/O) links exceeds multi-Gb/s, undesirable interferences from neighbor transmitting symbols and adjacent signal paths in electrical channel significantly hinder successful wired communications [1]. With rapidly enhancing transceiver capability with CMOS scaling, equalization techniques are required to mitigate inter-symbol interference (ISI) arising from signaling over band-limited channels. Meanwhile, high-density I/O applications have area and power restrictions, which limits complexity of transceiver architectures. This restriction obscures fully analog equalization due to resolution limitation [2] and difficulties to implement integrated analog delay cells [3]. Although a conventional binary receiver with multi-tap decision-feedback equalizer (DFE) helps further increasing achievable data rate (Fig. 1(a)) [4]-[6], analog blocks face harsher challenges due to process, voltage and temperature (PVT) variation. Besides, DFE itself is restricted to cancel post cursor ISI and less efficient for long tail ISI.

Increasing demands for advanced equalization and modulation techniques have led on-going interest in ADC-based high-speed links (Fig.1(b)) that leverage digital signal processing (DSP), which provides 1) robustness and adaptability in equalization, 2) flexibility for potential modulation schemes, 3) endurableness for PVT variations, and 4) scalability with CMOS technology improvement [7]-[11]. Despite the advantages of

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on mean-square error (MSE), such as signal-to-noise and distortion ratio (SNDR) or effective number of bits (ENOB) [9], [10], [12], [16]. To simplify analysis, ADC quantization distortion is often approximated as a normal distribution, and the total error variance is assumed to be the sum of the random noise and quantization distortion variance. While this is valid when random noise dominates over quantization distortion, this is not the case for low-resolution high-speed link systems.

The objectives of this thesis are to build statistical model for ADC-based link with limited quantization levels, explore performances and tradeoffs of existing solutions in modulation and equalization techniques for high-speed transceivers by statistical analysis, and propose a hybrid receiver which improves power efficiency of ADC-based links.

## **1.1 Organization of Thesis**

Chapter 2 presents an overview of statistical binary link modeling including band-limited channel, analog equalization, controlled signaling and non-ideal TX/RX timing. To extend previous statistical analysis for ADC-based link, statistical modeling for significant amount of quantization distortion by limited resolution ADC with following digital FFE-DFE is proposed.

Chapter 3 introduces an initial guidance in modulation selection among binary signaling (PAM-2), four-level pulse-amplitude-modulation (PAM-4) and duobinary in previous work and reviews the impacts of non-sufficient equalization, crosstalk and jitter on advance modulation selection by statistical analysis.

In Chapter 4, ADC resolution requirement reduction performed by PAE is reexamined by statistical modeling. Then, the new hybrid architecture which allows for additional saving of digital equalization power by the selective equalization technique is proposed. In addition, potential uses and drawbacks of embedded-DFE ADC is discussed.

Finally, Chapter 5 concludes the thesis with performance comparison for different modulation and equalization techniques and presents expected research extensions beyond this work.



## CHAPTER II

### STATISTICAL LINK MODELING

Statistical link modeling captures statistical natures of disturbances such as ISI, crosstalk, jitter and random noise, and estimates bit-error-rate (BER) in probability density function (PDF) domain. This better performs for fast and accurate link performance estimation with extremely low target BER, typically ranging from  $1e-12$  to  $1e-15$ , than the other approaches; transient simulation requires demanding tasks; worst case analysis is highly pessimistic [13]; mean-square-error (MSE) based analysis is based on Gaussian approximation for overall noise and distortion which is often invalid BER estimation for low target BER due to unbounded tail probability of a normal distribution [14]. This chapter summarizes statistical binary link modeling techniques in previous works and extends it for ADC-based link modeling.

#### **2.1 Binary Link Modeling**

Binary link consists of a channel, transmitter (TX), receiver (RX), analog equalizer, and timing circuits such as TX phase locked loop (PLL) and RX clock data recovery (CDR). This section presents an overview of statistical binary link modeling including the listed link components.

### 2.1.1 Channel Modeling

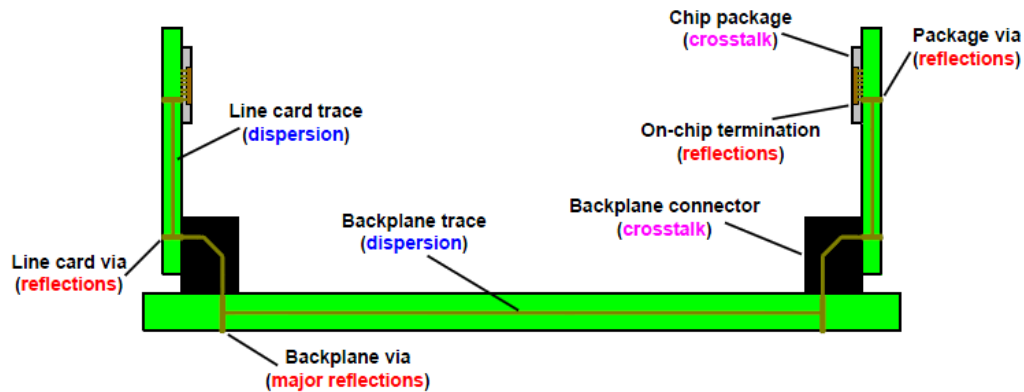


Fig. 2 Electrical backplane channel

Over a multi-Gb/s wired communications through electrical channels face three main disturbances causing signal interferences, which are dispersion, reflection and crosstalk [1]. Fig. 2 describes an electrical backplane channel components and sources of interferences. At first, dispersion occurs by limited high frequency current density at conductor surfaces (skin effect) and signal energy absorption into quickly rotating dielectric atoms along with alternating electric field due to high frequency signal (dielectric absorption). Secondly, discontinuity at vias and mismatch at on-chip termination reflect a part of transmitting signal, which affects on following signals. Thirdly, high frequency signal is leaked into neighbor channels through parasitics between densely placed and coupled wires. As a result, a high-speed data communication over an electrical channel suffers from band limitation due to dispersion, undesirable high frequency boosting due to reflections, and interferences from neighbor channels called crosstalk. A physical channel is often assumed as linear and time-

invariant (LTI) system and measured as a form of s-parameter, which stands for a ratio between input and output signals at different ports [17]. For instance,  $S_{21}$  in a two-port network presents the amount of output signal versus the input through the channel, while  $S_{11}$  shows the input reflection coefficient at given frequency components. Fig. 3 shows examples of the observed channel frequency responses of a communicating and crosstalk channels [18]. In addition to ISI and crosstalk, there exists another source increasing a chance of error, random noise. This is mainly caused by PVT variation and potential mismatches both in channel and TX/RX driver. Random noise is typically considered as an additive white Gaussian noise (AWGN) with zero mean and receiver's input referred sigma (standard deviation).

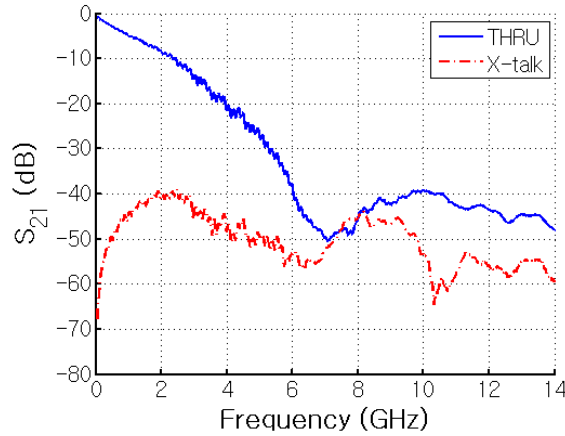


Fig. 3 Channel and crosstalk frequency responses [18]

Based on LTI assumption, a received pulse passed through a channel is obtained by convolving a unit pulse with channel impulse response which is inverse Fourier transform of the channel frequency response [19]. Channel pulse response in Fig. 4(a)

clearly shows how adjacent symbols interfere with each other. Given '101' of the transmitting data pattern, parts of pre and post '1' symbols appears at '0' symbol reception and causes an error as shown in Fig. 4(b). Moreover, sampled values from adjacent crosstalk channel (Fig. 5) and random noise disturb transmitted symbol detection.

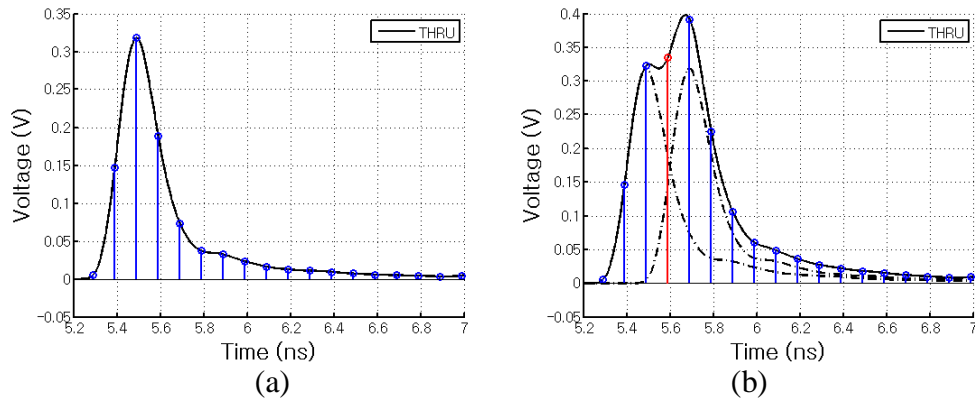


Fig. 4 (a) Channel pulse response and its sampled response and (b) {101} pattern pulse response and its sampled response at 10Gb/s in the thru channel in Fig. 3.

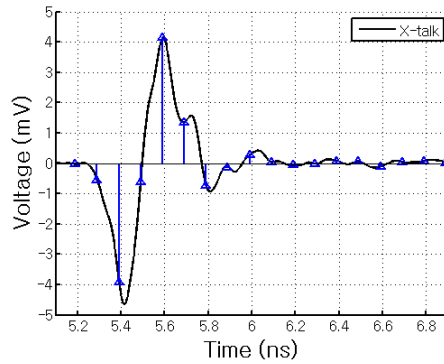


Fig. 5 Crosstalk channel pulse response and its sampled response at 10Gb/s in the thru channel in Fig. 3.

PDF of overall ISI, crosstalk and random noise is generated by convolving each independent interfering symbol's PDF [15]. A received and sampled signal is described as,

$$y_k = h_0 I_k + \sum_{n \neq k} h_{n-k} I_k + \sum_{\forall m} g_{m-k} \tilde{I}_m + Z_k, \quad (2.1)$$

where  $k$  is cursor indicator,  $h$  and  $g$  are sampled pulse responses of through and crosstalk channels,  $I$  and  $\tilde{I}$  are transmitted symbols through corresponding channels, and  $Z$  is AWGN. A transmitting symbol probability mass function (PMF) is known, for example, PAM-2 signaling has two equiprobable symbols,  $\{-1, 1\}$ . If transmitting symbols are independent (not coded), the total ISI and crosstalk is sum of the independent symbols scaled by channels. Since a PDF of the sum of the independent random variables is same as convolution of all individual independent PDFs, total noise and distortion PDF and its cumulative distribution function (CDF) are obtained by convolving all ISI PMFs and random noise PDF. A voltage margin referred to a certain target BER (statistical eye opening) is typically presented by bathtub curve, which constructed by two flipped CDFs whose means are the corresponding cursor amplitudes. Fig. 6 shows the procedure to create a bathtub curve with statistical modeling.

Statistical modeling requires more computational complexity than MSE-based analysis, but is valued in accurate low BER estimation. To simplify BER estimation, overall noise and distortion distribution is assumed as a normal distribution and error probability is evaluated based on signal-to-noise-and-distortion ratio (SNDR) [19].

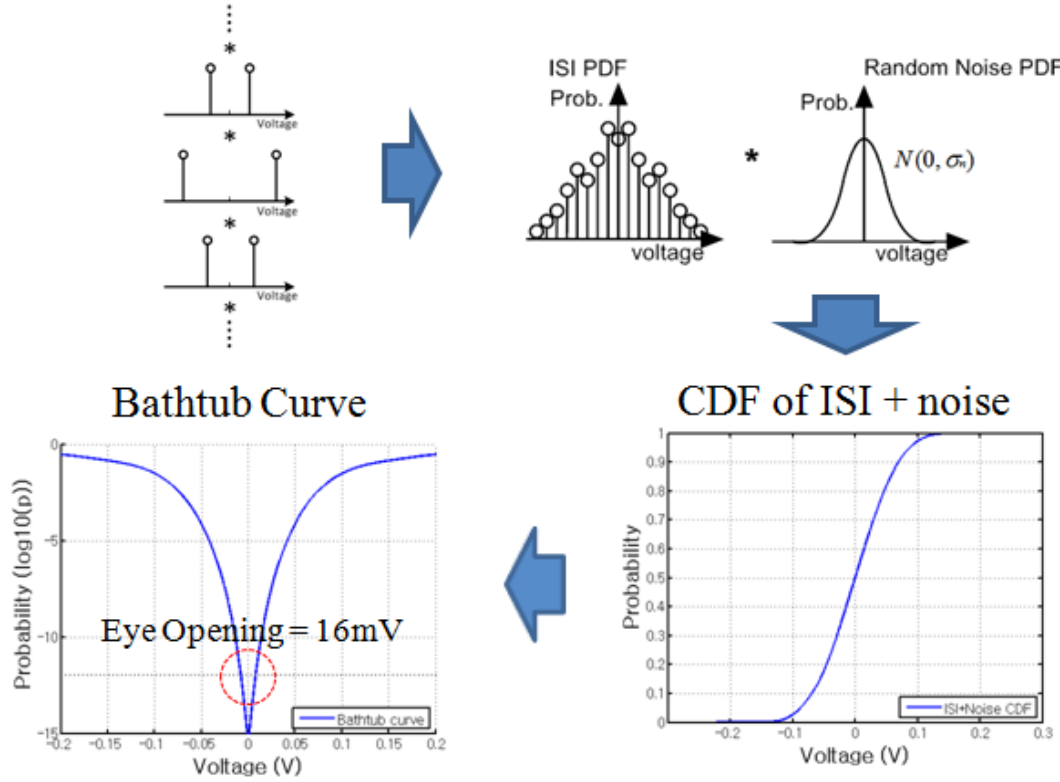


Fig. 6 Procedure to create ISI + random noise CDF and estimate statistical eye opening with PAM-2 signaling and  $1e-12$  of the target BER

$$SNDR = \frac{P_{signal}}{P_{noise} + P_{distortion}}, \text{ and} \quad (2.2)$$

$$P_e \approx Q \left( \sqrt{\frac{d_{min}^2/4}{\sigma_{noise}^2 + \sigma_{distortion}^2}} \right) = Q(\sqrt{SNDR} - 1) \quad (2.3)$$

For instance, in case of (1) with PAM-2 signaling,

$$P_{e,PAM2} \approx Q \left( \sqrt{\frac{h_0^2}{\sum_{n \neq 0} h_n^2 + \sum_{m \neq 0} g_m^2 + \sigma_z^2}} \right) \quad (2.4)$$

where  $Q$  is the complementary function defined as

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp\left(-\frac{u^2}{2}\right) du, \quad (2.5)$$

which is based on CDF of the standard normal distribution. Fig. 7 presents comparison between the CDFs of ISI and random noise by statistical modeling, and approximated normal distribution whose variance is same as the former. This shows that the MSE based analysis is extremely pessimistic at low BER range which potentially results in over-design, and invalid for qualified BER estimation with low target BER due to bound of ISI and crosstalk distribution [14]. A statistical BER estimator improves estimated BER accuracy by replacing the complementary function to bounded CDF relying on the statistical modeling.

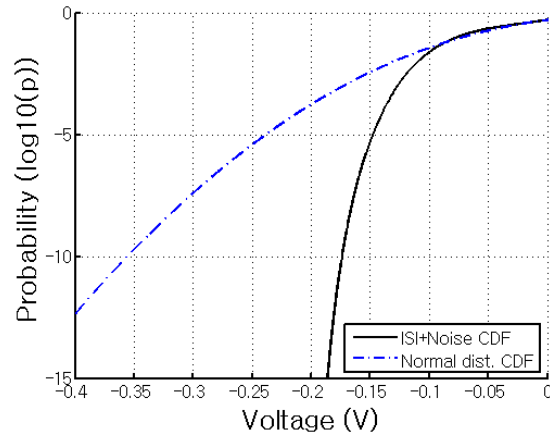


Fig. 7. Comparison between CDFs by statistical modeling and approximated normal distribution

### 2.1.2 Equalization and Modulation Modeling

Feed-forward equalizer (FFE), continuous-time linear equalizer (CTLE) and DFE are the most common in binary high-speed link [4]-[6], [20]-[21]. To avoid implementing analog delay cell, FFE is typically implemented at transmitter, so called pre-emphasis. CTLE compensates high frequency loss by controlled in-band zero and poles. DFE subtracts the exact amount post cursor ISI induced by post symbols based on previous decisions. Recently, a RX DFE with infinite impulse response (IIR) filter for second and further post cursor ISI has been studied to cancel long tail pulse response other than DFE taps [22]-[23].

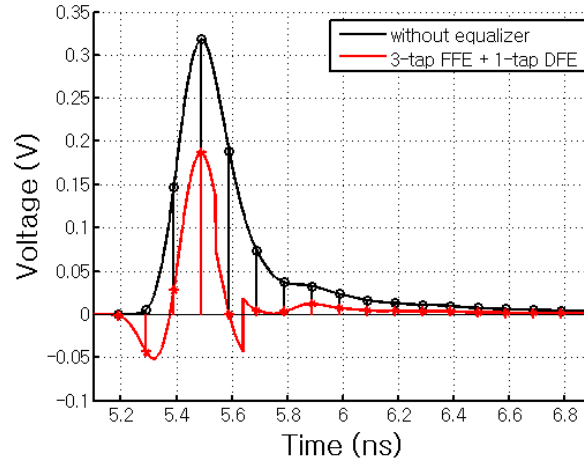


Fig. 8 Equalized pulse response by 3-tap FFE and 1-tap DFE

LTI assumption allows for applying any type of linear equalizer, such as FFE, CTLE, either in frequency domain by multiplying equalizer frequency response to channel frequency response or in time domain by convolving equalizer impulse response



with channel pulse response. Although DFE is non-linear, ideal functionality without error propagation due to potential wrong pre-decisions is assumed when achieving low target BER. Ideal DFE always subtracts the exact amplitude of corresponding post cursor ISIs for one unit interval (UI) per each DFE tap. Fig. 8 presents the pulse responses before and after equalization with FFE-DFE. Statistical modeling of residual ISI after equalization follows the same procedures discussed in previous section with the equalized pulse response.

Given channel information, FFE coefficients are typically optimized based on minimum-mean-square-error (MMSE) criteria [19]. Zero forcing FFE has MMSE of ISI when its coefficients are

$$C_{ZF-FFE} = (H^T H)^{-1} H^T Y_{desired}, \text{ where} \quad (2.6)$$

$$H = \begin{bmatrix} h(0) & 0 & 0 & 0 & 0 \\ h(1) & h(0) & 0 & 0 & 0 \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & 0 & h(k-1) & h(k-2) \\ 0 & 0 & 0 & 0 & h(k-1) \end{bmatrix}, \quad C_{ZF-FFE} = \begin{bmatrix} c(0) \\ c(1) \\ \vdots \\ c(l-1) \end{bmatrix},$$

$h$  is  $k$ -taps sampled channel pulse response,  $c$  is FFE coefficients with  $l$ -taps and  $Y_{desired}$  is the desired output, such as the vector having 1 for the cursor tap and 0 for all the others (for PAM modulation). Since FFE scales AWGN either, considerable random noise should be taken into account. With AWGN, MMSE-FFE coefficients are optimized when

$$C_{MMSE-FFE} = (H^T H + N)^{-1} H^T Y_{desired}, \quad (2.7)$$

where  $N$  is  $l$  by  $l$  identical matrix multiplied by variance of independent noise. Furthermore, with ideal DFE assumption, MMSE-DFE coefficients [24] are optimized when

$$W_{MMSE-DFE} = (\tilde{H}^T \tilde{H} + \tilde{N})^{-1} \tilde{H}^T Y_{desired}, \text{ where} \quad (2.8)$$

$$\tilde{H} = [H | I_{DFE}], \quad W_{MMSE-DFE} = \begin{bmatrix} c(0) \\ c(1) \\ \vdots \\ c(l-1) \\ b(0) \\ \vdots \\ b(m-1) \end{bmatrix}, \quad \tilde{N} = \begin{bmatrix} N & 0 \\ 0 & 0 \end{bmatrix}$$

$c$  and  $b$  are  $l$ -taps FFE and  $m$ -taps DFE coefficients, respectively, and  $I_{DFE}$  is the matrix whose components are

$$I_{DFE}(k, j) = \begin{cases} 1, & \text{if } k = k_{cursor} + j \\ 0, & \text{otherwise.} \end{cases}$$

Note that MMSE does not necessarily results in optimal BER when ISI is not controlled enough, since the ISI PDF differs from a normal distribution PDF as mentioned. However, it offers useful initial guess for finding optimal equalizer coefficients, and further optimization can be done by coefficient sweeping in statistical analysis.

Pulse amplitude modulation (PAM) is easily modeled by different transmitting symbol PDF. For instance, PAM-4 has four symbols,  $\{-1 -1/3 1/3 1\}$  and each ISI tap has four equiprobable values. Thus, an independent transmitting symbol has four-value PMF instead of two-value PMF in PAM-2. Equalizer coefficient optimization procedure is exactly same for any types of PAM. There exist  $M-1$  thresholds for PAM- $M$  and pairs of thresholds from outer two to inner two excepting zero are symmetric. Ideal thresholds

places exactly at the middle of two adjacent constellations, statistical eye opening are the same for all  $M-1$  thresholds. In practice, thresholds cannot be adapted to non-ideal sampling point, BER at outer threshold is typically worse than inner one with given fixed thresholds.

Since duobinary modulator includes a channel [25], MMSE equalizer coefficients are obtained by simply changing the desired vector to have two cursors whose values are 0.5 and 0.5 instead of 1 for PAM. Transmitted data is same as PAM-2 but the received signal is

$$y_k = h_0 I_k + h_1 I_{k+1} + \sum_{n \neq k, k+1} h_{n-k} I_k + \sum_{\forall m} g_{m-k} \tilde{I}_m + Z_k, \quad (2.9)$$

where  $k$  and  $k+1$  are cursor and the first post cursor,  $h$  and  $g$  are sampled pulse responses of through and crosstalk channels,  $I$  and  $\tilde{I}$  are transmitted binary symbols,  $\{1, -1\}$ , and  $Z$  is AWGN. Although duobinary has three possible received symbol,  $\{\pm(h_0 + h_1), 0\}$ , mismatch between cursor and the first post cursor separates zero symbol into two and the distance between constellations is degraded by  $|h_0 - h_1|$ .

### 2.1.3 Jitter Modeling

While statistical modeling discussed in section 2.1.1-2 is sample-based, transmitting and sampling timing is not ideal because of jitter at PLL and CDR. Jitter analysis has been mature with linear modeling of PLL and CDR [14], [26]-[29]. Although phase error of the TX/RX clocks from diverse sources causing jitter in PLL and CDR is analyzable by jitter transfer function, linearized jitter model often provides

too much details with extreme complexity because of various sources of jitter in clocking circuits that are not necessarily Gaussian distribution [14], and correlation between TX jitter and RX jitter which is also disturbed by channel ISI, crosstalk and even DFE subtraction [13]. To simplify statistical analysis, RX jitter is often modeled by dual-Dirac jitter that approximates jitter distribution as combination of deterministic jitter (DJ) and random jitter (RJ) [30], and TX jitter is modeled by combination of periodic and random jitter [14], [31].

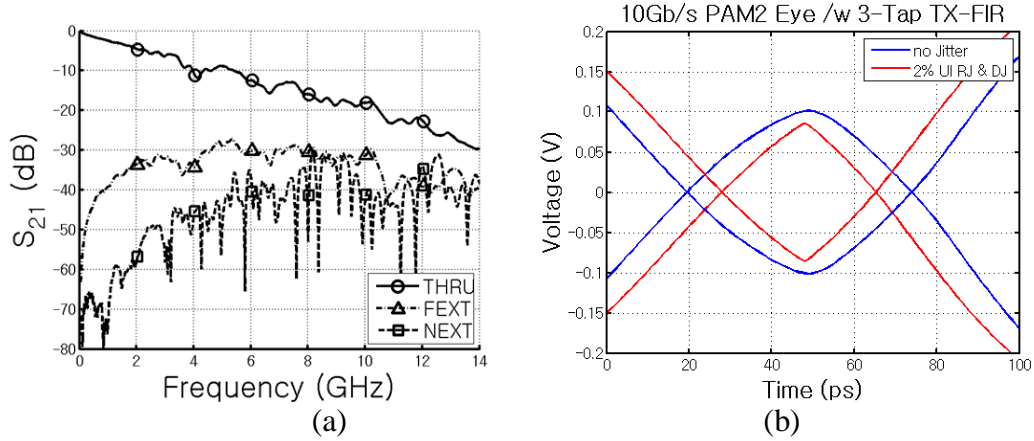


Fig. 9 (a) Thru and crosstalk channel frequency responses [18] and (b) statistical eye degradation by 2%UI DJ and 2%UIrms RJ of RX jitter

RX jitter distribution in time is transferred to PDF in voltage domain by a family of PDFs obtained at different sampling timing. Since

$$P_{V,T}(v, t) = P_{V|T}(v|t)P_T(t), \quad (2.10)$$

sum of conditional PDF at the given sampling time multiplied by probability to have that sampling time is the PDF with RX jitter. Fig. 9(b) presents statistical eye degradation by

2%UI of DJ and 2%UI of RJ sigma, which is simulated with the through channel with two crosstalk channels in Fig. 9(a). The simulation setting are 10Gb/s of data rate, 3-tap FFE with PAM-2 signaling and 1mVrms random jitter. In Fig. 10, the procedure of statistical RX jitter modeling and the result is compared to 1k-bit transient simulation with through channel.

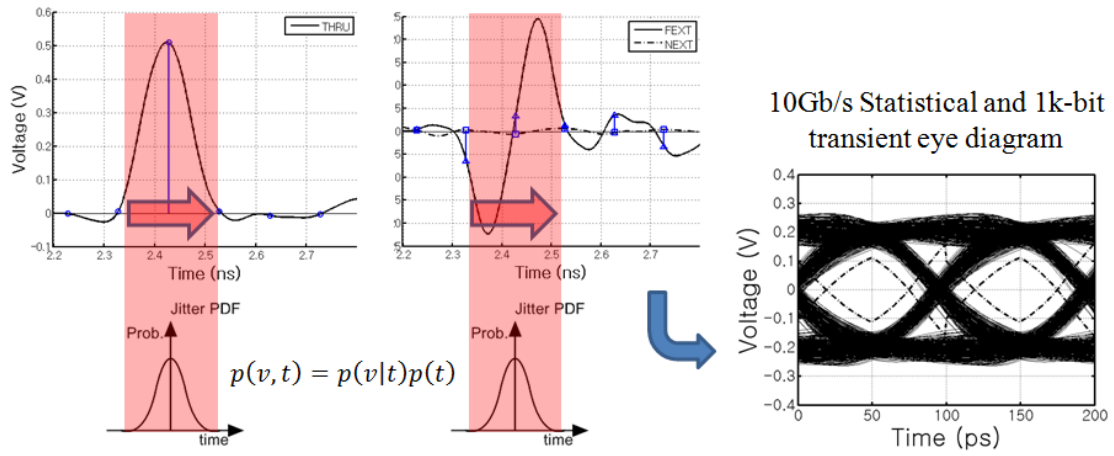


Fig. 10 Procedure of RX jitter modeling and comparison between statistical eye contour referred to 1e-12 of BER and 1k-bit transient simulation eye diagram

TX jitter causes varying TX pulse widths that results in changes of the channel pulse response. As a result, this does not only affects on RX jitter but also generates received signal amplitude variation. If TX jitter is totally random and limited to small variance, random TX jitter is modeled by jitter transfer function from jitter impulse response. Independent random TX jitter occurs at both of rising and falling edges and RX signal voltage varies linearly if the TX timing variance is small enough. Thus, impulse TX timing variance is transferred to jitter sensitivity function (JSF) [14] in

voltage as described in Fig. 11. Those two JSF at rising and falling edges are highly correlated such that increasing pulse width of current symbol directly results in decreasing pulse width of next symbol, so that

$$\sigma_{v,RX}^2 = \sum_{\forall n} \{JSF_{falling}(n) + JSF_{rising}(n)\}^2 \sigma_{t,TX}^2. \quad (2.11)$$

where  $\sigma_{t,TX}^2$  is the variance of a normal TX timing distribution and  $\sigma_{v,RX}^2$  is the corresponding voltage variance of the normal distribution in voltage.

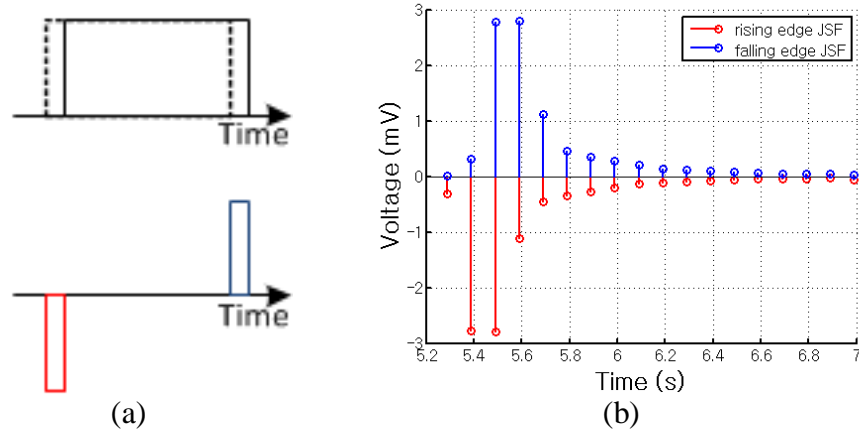


Fig. 11 (a) TX impulse jitter model and (b) jitter sensitivity function (JSF) at the channel output

Periodic TX jitter should be considered in a sequence of varying pulse widths [31]. For instance, duty cycle jitter, the simplest periodic jitter, is modeled by combining two pulse responses originated from longer pulse width and shorter one which are transmitted one-by-one in a sequence. Both cases, whether cursor is generated by longer or shorter pulses, are equiprobable, so that two statistical PDF from each case are

generated and averaged as shown in Fig. 12. This method, however, still requires demanding tasks as increasing complexity of periodic sequences [13]. Nevertheless, considering the purpose of statistical analysis, that is providing the qualified guidance for system-level transceiver design, this modeling is useful by applying proper marginal conditions with simplified jitter models.

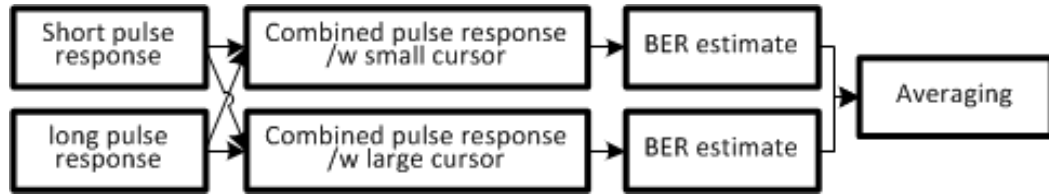


Fig. 12 Procedure for duty cycle TX jitter modeling

## 2.2 Analog-to-Digital Converter (ADC) Based Link Modeling

Quantization distortion of a uniform ADC is typically modeled as an additive white uniform distribution whose variance is

$$\sigma_{qz}^2 = E[(X - \mu)^2] = \int_{-\frac{\Delta_{qz}}{2}}^{\frac{\Delta_{qz}}{2}} \frac{1}{\Delta_{qz}} x^2 dx = \frac{\Delta_{qz}^2}{12}, \quad (2.12)$$

where  $\Delta_{qz}$  is the distance between quantization levels. To simplify analysis, quantization distortion is often approximated as a normal distribution [9], [10], [12], and BER is estimated as follows.

$$SNDR = \frac{P_{signal}}{P_{noise} + P_{distortion} + P_{quantization}}, \text{ and} \quad (2.13)$$

$$\begin{aligned}
P_{e,PAM2} &\approx Q\left(\sqrt{\frac{d_{min}^2/4}{\sigma_{noise}^2 + \sigma_{distortion}^2 + \sigma_{quantization}^2}}\right) \\
&= Q\left(\sqrt{\frac{h_0^2}{\sum_{n \neq 0} h_n^2 + \sum_{\forall m} g_m^2 + \sigma_z^2 + \Delta_{qz}^2/12}}\right) \\
&= Q(\sqrt{SNDR - 1}).
\end{aligned} \tag{2.14}$$

Also, ADC resolution requirements at a given target BER is analyzed relying on the definition of effective number of bits (ENOB) [16], that is

$$ENOB = \frac{SNDR - 1.76}{6.02}. \tag{2.15}$$

Although this is valid when random noise is dominant so that overall noise and quantization distortion distributes as the closed form of the normal distribution, typical ADC-based link with limited ADC resolutions to 4-6 bits is not the case.

Previously, in order to improve accuracy for systems with limited quantization levels, normal and uniform distributions were combined to derive BER for M-PAM with fading [32]. However, this approach is restricted to analyze the output of the ADC. When the quantized received signal passes through a following digital FFE, the quantization error is scaled by FFE coefficients and scaled error distributions at different sampling times are added up. Since a sum of non-identical uniform distributions has discontinuous points on its PDF, analytical BER estimation is difficult. This issue is also discussed in [7], in which it is argued that required noise sigma to achieve  $BER < 1e-17$  is 15~25% of ADC LSB for 3.5bit ADC based on empirical observations, while this is not a general solution.



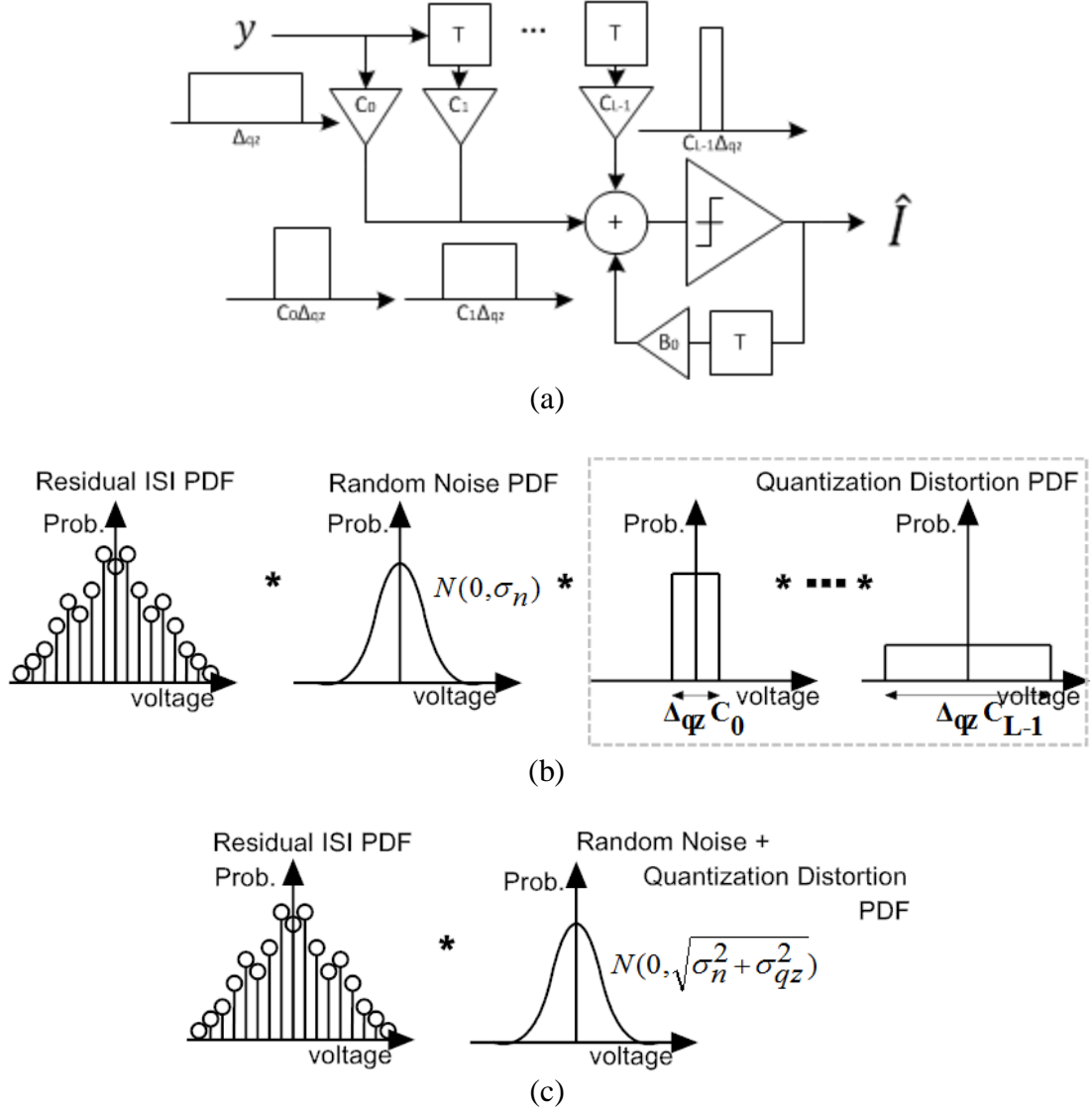


Fig. 13 (a) Quantization distortion in digital FFE-DFE, (b) proposed statistical modeling for ADC-based receiver with digital FFE-DFE, and (c) comparing statistical modeling with Gaussian-approximated quantization distortion

In the specific case of digital equalization, digital FFE-DFE (Fig. 13(a)) which is a common equalizer for low-complexity and power efficient applications, statistical modeling allows for link performance evaluation without Gaussian approximation. The

proposed statistical analysis utilizes the uniformly distributed quantization noise PDFs scaled by the FFE coefficients for improved BER estimation accuracy with the low ADC resolutions common in high-speed link systems. A PDF of the sum of the independent and uniformly distributed random variables is obtained by convolving all the individual PDFs. This allows for the channel and analog front-end modeling discussed in Section 2.1 to be expanded to the modeling of ADC-based links, as shown in Fig. 13(b). Note that DFE, which mitigates residual ISI by eliminating corresponding post cursor ISI taps, does not affect quantization distortion.

The improved BER estimation accuracy by the proposed technique of scaling quantization error with the FFE coefficients is shown in the bathtub curves at different ADC resolution settings and FFE tap numbers in Fig. 15-16. For the fair comparison of the improvement by the proposed method, comparing bathtub curve is generated with statistical modeling for ISI and Gaussian approximation for quantization distortion (Fig. 13(c)), and digital FFE coefficients are normalized not to scale a normal distribution. Here, the two quantization distortion modeling techniques of Fig. 13 and also transient simulation results are compared for a 10Gb/s data rate on two different backplane channels in Fig. 14. 4/1-tap and 9/1-tap digital FFE/DFEs are used for the channel 1 and 2, respectively. Other simulation settings include the use of PAM-2 signaling,  $1\text{mV}_{\text{rms}}$  AWGN,  $1\text{V}_{\text{pp}}$  input dynamic range, and 8-bit digital equalizer resolution. To observe the statistical bound,  $1\text{E}9$  bits are simulated in transient. For this, all three cases are sample-based simulation, which means the jitter is not considered. However, since jitter in statistical analysis is applied by combining PDFs built with different transmitting pulse

width and sampling time, the verified accuracy in sample-based simulation confirms the accuracy with jitter. While the interest is the results with 4-6 bits of ADC resolution, 8-bit ADC is also simulated for sanity check, in which all the three compared cases expect to be matched.

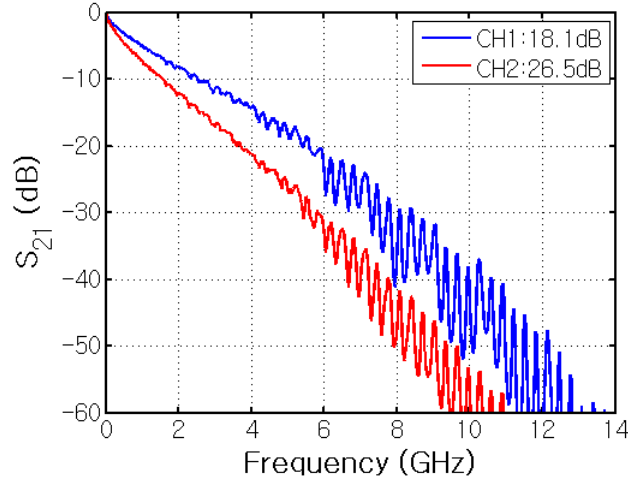


Fig. 14 Channel frequency responses [18]

As shown in Fig. 15(b)-(d) and Fig. 16(b)-(d), utilizing the proposed quantization distortion modeling technique allows for improved tracking of transient simulation results for the typical 4-6 bit ADC resolution settings for high-speed link applications. Note that when quantization distortion is small, as in Fig. 15(a) and Fig. 16(a) with 8-bit ADC resolution, both techniques track transient simulation results well. Significance is Gaussian approximation is often pessimistic and fails to detect large eye opening in Fig. 15(d) that potentially results in over-design. Interestingly, Fig. 16(d) shows possibility of system fail due to optimistic BER estimation in Gaussian approximation. This is due to

the bounded convolution of uniform distributions widening by the absolute sum of the FFE coefficients, while the variance of the quantization distortion depends on the square sum of the coefficients. Thus, as the number of FFE taps increases, the magnitude of the FFE cursor tap increasingly spreads into pre- and post-cursor taps and widens the quantization distortion bounds.

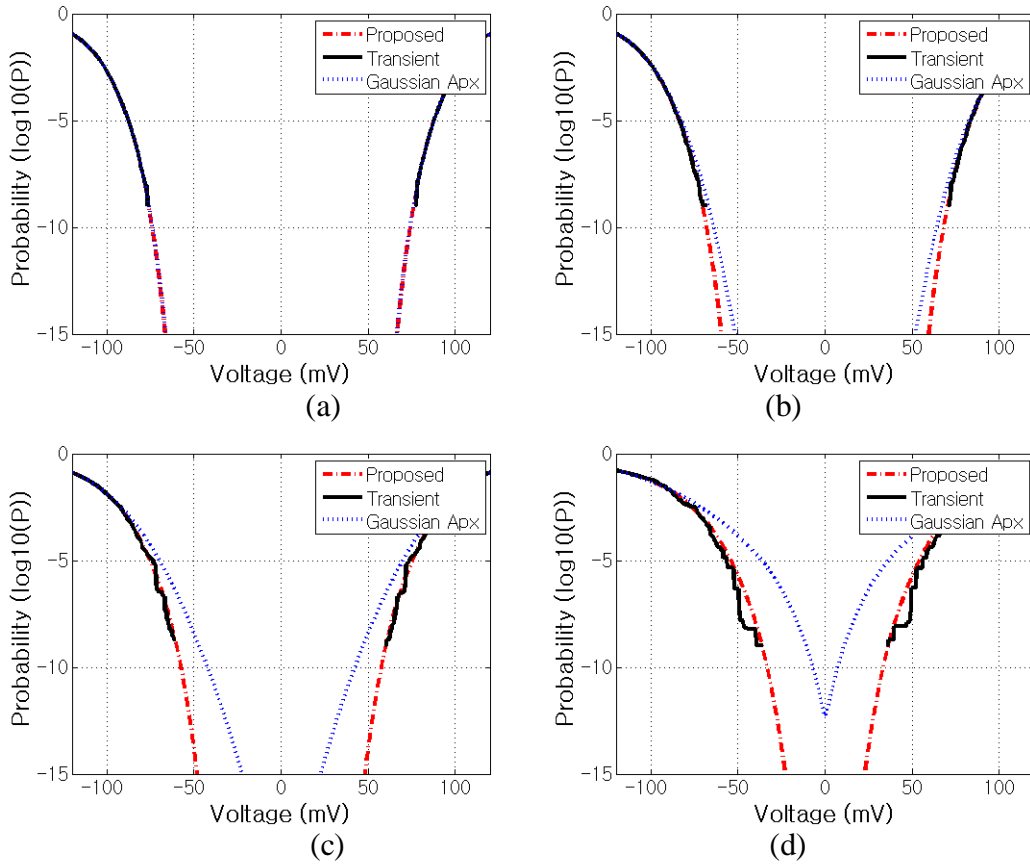


Fig. 15 Bathtub curve comparison at channel 1 in Fig. 14 (a) with 8-bit ADC, (b) 6-bit ADC, (c) 5-bit ADC, and (d) 4-bit ADC

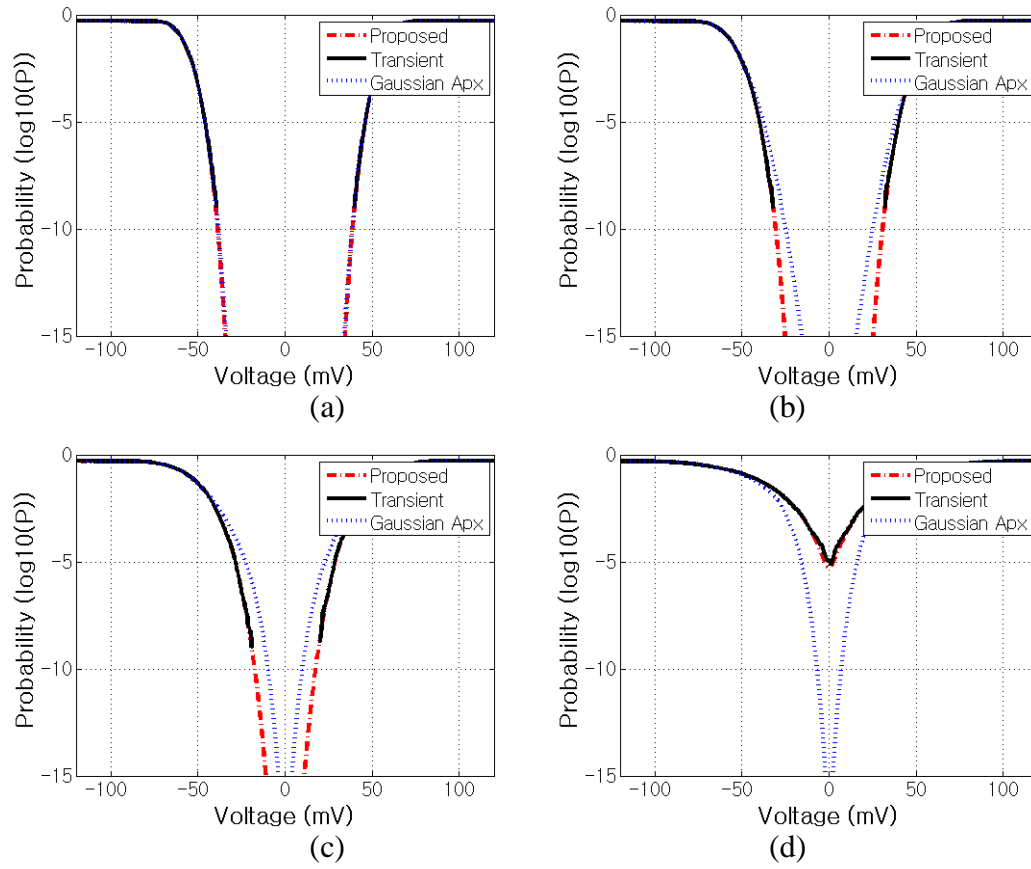


Fig. 16 Bathtub curve comparison at channel 2 in Fig. 14 (a) with 8-bit ADC, (b) 6-bit ADC, (c) 5-bit ADC, and (d) 4-bit ADC

## CHAPTER III

### MODULATION TECHNIQUES AND ANALYSIS

In addition to equalization, modulated signaling allows for controlling ISI over band-limited channel. Besides, an ADC-based receiver is flexible for different types of signaling that encourages uses of optimal combinations of equalization and modulation techniques. Due to limited transceiver complexity by power and area restriction of compact high-speed I/O applications, a couple of modulation techniques, PAM-4 and duobinary, have been used in wired communications [25], [29], [33]. In this chapter, PAM-2, PAM-4 and duobinary are compared by statistical analysis and anticipate the possible solutions for future ADC-based links.

#### **3.1 Initial Guidance for Modulation Selection**

Modulation selection is the critical issue because finding the best performer among different modulation schemes depends on not only channel loss but also the distance between signal constellations. While each of PAM-4 and duobinary performances is compared to PAM-2 in [29], [33], the comparison among three of them is proposed in [25] with a simple and initial guidance based on effective Nyquist frequency channel loss. Due to peak power limitation of transmitting signal, maximum boosting at Nyquist frequency is restricted to 0dB, that is, channel loss at Nyquist frequency cannot be overcome. Given data rate, PAM-4 has half of symbol rate or half effective Nyquist frequency compared to PAM-2. Thus, even though ISI is equalized

enough for both of cases, PAM-2 suffers from higher channel loss. Controlled ISI by duobinary provides two third of effective Nyquist frequency reduction.

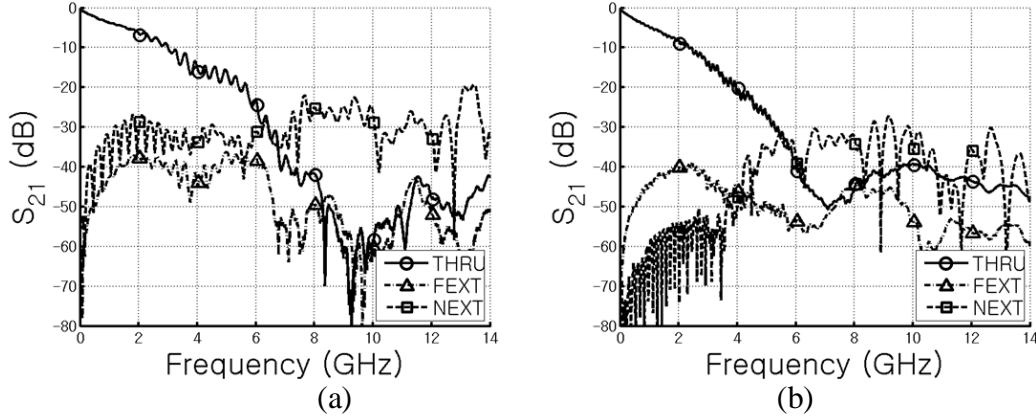


Fig. 17 Through and crosstalk channels [18] for (a) case 2 and (b) case 3

However, since PAM-4 and duobinary have four levels and three levels of signal constellation within given input dynamic, the minimum distances between constellations are limited by one third and half, respectively. With the assumption of enough equalization, modulation gain is calculated by subtracting constellation loss from the gain obtained by relaxed channel loss. For instance, three cases at different channels and data rates are carefully selected such that each of the cases results in different modulation selection; case 1 is 10Gb/s at channel in Fig. 9(a); case 2 is 10Gb/s at channel in Fig. 17(a); case 3 is 8Gb/s at channel in fig. 17(b). Channel losses at effective Nyquist frequencies and constellation losses are profiled in Table I. Based on the loss profiles, modulation gain is summarized in Table II. For example, when PAM-4 is used instead of PAM-2 in case 1, 3.1dB of loss reduction is achieved while the distance

between constellations is reduced by 9.54dB. As a result, total 6.44dB loss is expected with PAM-4 selection versus PAM-2. Finally, the initial guidance informs that PAM2 performs the best for case 1, PAM4 for case 2, and duobinary for case 3.

Table I Channel loss profiles

Loss @ effective Nyquist freq	PAM-2	Duobinary	PAM-4
Case 1	8.9dB	6.0dB	5.8dB
Case 2	18.2dB	12.6dB	7.9dB
Case 3	21.3dB	11.4dB	8.2dB
Constellation loss	0dB	6dB	9.54dB

Table II Modulation gain

Modulation gain	Duobinary vs. PAM-2	PAM-4 vs. PAM-2	Selection
Case 1	-3.10dB	-6.44dB	PAM-2
Case 2	-0.40dB	0.76dB	PAM-4
Case 3	3.90dB	0.36dB	Duobinary

### 3.2 Advanced Modulation Selection

The simple principle for modulation selection discussed in previous section does not consider insufficient equalization and effect of crosstalk and jitter. In [29], effects of non-ideal transition of TX pulse, jitter and different equalization complexity on PAM-2 and duobinary performance are compared by statistical modeling, but crosstalk is not considered. Moreover, improved performance comparison between PAM-4 and



duobinary versus PAM-2 is very important to predict future trends in high-speed link design. From the initial guidance, we know that channel loss range for selecting duobinary places between those for PAM-2 and PAM-4. In this work, the three modulation techniques are compared by statistical analysis.

Table III Simulation setting for the three cases in Table I

Input dynamic	1Vppk
Random noise	1mVrms
RX DJ	1%UI
RX RJ	1%UIrms
Target BER	1E-12
Crosstalk	FEXT, NEXT
Equalizer	3-tap FFE
Equalizer resolution	Ideal

Here, the three cases selected in Table I are reexamined by statistical analysis. Simulation settings are summarized in Table III and the results are presented in Table IV-VI, as well as the optimized coefficients of 3-tap FFE by MMSE criteria. Also, from the statistical eye opening, modulation gain is recalculated and compared in table VII. As expected, the selected modulation scheme for each case has the largest eye opening. However, performance differences are not well matched to the estimated modulation gain by the simple principle presented in Section 3.1. For example, the initial guidance shows that the expected eye openings for the three modulation schemes in case 3 are

similar, but PAM-4 shows remarkable improvement while duobinary has worse performance than expected. This means, there exists possibility that the selected modulation scheme is changed by non-ideality of channel and equalizer capability.

Table IV Case 1: FFE coefficients and statistical eye opening

	$a_{-1}$	$a_0$	$a_1$	1e-12 stat. eye H(mV)	1e-12 stat. eye W(ps)
PAM2	-0.0492	0.7177	-0.2331	220.4	56
PAM4	-0.0179	0.8824	-0.0997	117.8	80
DUO	0.4951	0.3273	-0.1776	154.7	57

Table V Case 2: FFE coefficients and statistical eye opening

	$a_{-1}$	$a_0$	$a_1$	1e-12 stat. eye H(mV)	1e-12 stat. eye W(ps)
PAM2	-0.1669	0.5994	-0.2337	14.2	13
PAM4	-0.0470	0.7972	-0.1559	44.4	36
DUO	0.7246	-0.2669	0.0086	8.3	7

Table VI Case 3: FFE coefficients and statistical eye opening

	$a_{-1}$	$a_0$	$a_1$	12-12 stat. eye H(mV)	1e-12 stat. eye W(ps)
PAM2	-0.1685	0.5917	-0.2398	54.2	41.25
PAM4	-0.0459	0.7767	-0.1774	58.4	65
DUO	0.7302	-0.2297	-0.0401	62	47.5

Table VII Modulation gain by statistical analysis

Modulation gain	Duobinary vs. PAM-2	PAM-4 vs. PAM-2	Selection
Case 1	-3.07dB	-5.44dB	PAM-2
Case 2	-4.66dB	9.90dB	PAM-4
Case 3	1.17dB	0.65dB	Duobinary

Since equalized frequency response by ideal duobinary has a null at Nyquist frequency ( $1/2T_b$ ) as shown in Fig. 18, a band-limited channel is typically included into the duobinary modulator, which relaxes equalization requirement [25]. Fig. 19 presents maximum achievable data rate with different complexity of FFE at channel in case 3. As the number of FFE taps increases, the available data rate also increases in PAM-2 and PAM-4. However, this trend is not shown in duobinary. Duobinary achieves the highest data rate with 2-tap FFE due to relaxed equalization requirement, while there is not notable improvement with increasing FFE taps more than two.

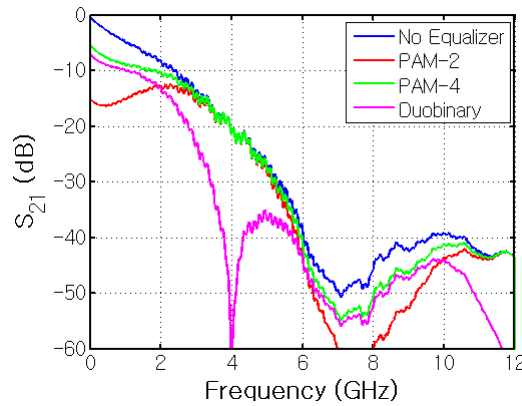


Fig. 18 Pre and post equalized channel frequency responses by 3-tap FFE with PAM-2, PAM-4 and duobinary

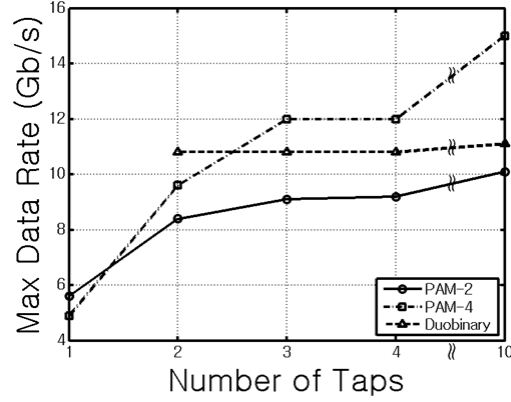


Fig. 19 Maximum achievable data rate with different number of FFE taps

In addition to limited duobinary transition to the adjacent level, prohibited patterns in duobinary,  $\{101\}$  and  $\{-10-1\}$ , provide larger voltage margin around ideal sampling point. However, the duobinary cursor, a combination of two cursors, suffers from mismatches due to non-ideal sampling timing. Thus, duobinary eye opening is more quickly degraded than PAM-2 as jitter increases. Clearly, PAM-4 has two times larger UI than the others, so that it has the best jitter tolerance. Fig. 20 presents the eye degradation due to increasing random jitter sigma in case 3. As a result, increasing jitter reduces the duobinary selection range versus PAM-2 and PAM-4.

Finally, crosstalk tolerances in different modulation techniques are analyzed by statistical modeling. In [33], it is argued that limited transitions to adjacent levels in duobinary improve immunity to crosstalk and reflection comparing to PAM-2 and PAM-4. This fact is shown in crosstalk frequency response filtered by FFE in duobinary (Fig.

21). However, PAM-4 has diminished average transmitting bit energy by factor of 5/9 since

$$E_{b,avg,PAM4} = \frac{1}{4} \left\{ (I)^2 + \left(\frac{1}{3}I\right)^2 + \left(-\frac{1}{3}I\right)^2 + (-I)^2 \right\} = \frac{5}{9} E_{b,avg,PAM2}. \quad (3.1)$$

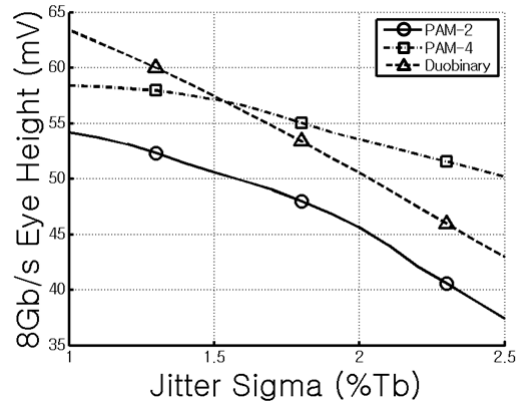


Fig. 20 Eye degradation by increasing RX random jitter sigma in case 3

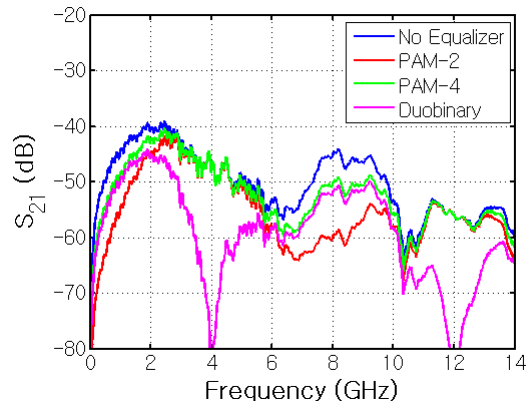


Fig. 21 Crosstalk channel frequency responses filtered by 3-tap FFE in case 3

Consequently, the variance of ISI and crosstalk in PAM-4 is

$$\sigma_{\text{PAM4}}^2 = \frac{5}{9} \sum_{i \neq k}^N h_{i,\text{PAM4}}^2 + \frac{5}{9} \sum_i^N \sum_j^M g_{ij,\text{PAM4}}^2 \quad (3.2)$$

where  $N$  is the channel length,  $M$  is the number of crosstalk channels,  $h_{i,\text{PAM4}}$  are the equalized and sampled thru channel pulse response and  $g_{ij,\text{PAM4}}$  are the  $j$ -th channel's sampled crosstalk pulse responses filtered by an equalizer, while those of PAM-2 and duobinary are

$$\sigma_{\text{PAM2}}^2 = \sum_{i \neq k}^N h_{i,\text{PAM2}}^2 + \sum_i^N \sum_j^M g_{ij,\text{PAM2}}^2, \text{ and} \quad (3.3)$$

$$\sigma_{\text{DUO}}^2 = (|h_{k,\text{DUO}}| - |h_{k-1,\text{DUO}}|)^2 + \sum_{i \neq k, k-1}^N h_{i,\text{DUO}}^2 + \sum_i^N \sum_j^M g_{ij,\text{DUO}}^2, \quad (3.4)$$

where the first term in duobinary presents the cursor mismatch. Thus, crosstalk tolerance capability depends not only on the amount of tolerated crosstalk and reflection but also the reduced bit energy passing through a crosstalk channel. The impact of crosstalk on performances of different modulation schemes is presented in Fig. 22. In addition to the previous simulation result in case 3, two more cases without crosstalk and with 6dB boosted crosstalk are compared. The result shows that selected signaling is reversed to PAM-4 with boosted crosstalk, while eye opening further increases without crosstalk.

In summary, based on the initial guidance, PAM-2 is selected for low loss channel, duobinary for mid loss channel, and PAM-4 for high loss channel. However, duobinary selection range is widened as equalization capability is limited, while increasing jitter and crosstalk get narrow that range as shown in Fig. 23. Considering

that ADC-based link targets high low channel requiring larger number of FFE to control significant amount of pre-cursor ISI, jitter and crosstalk become severe as increasing data rate while enough equalization is applied in DSP. Therefore, PAM-4 is more promising to use in ADC-based link than duobinary.

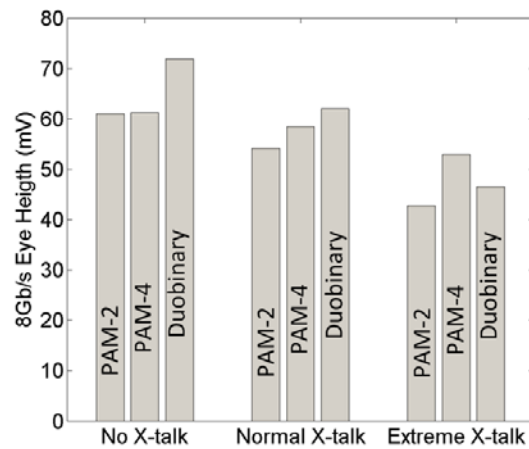


Fig. 22 Eye degradation by increasing crosstalk in case 3

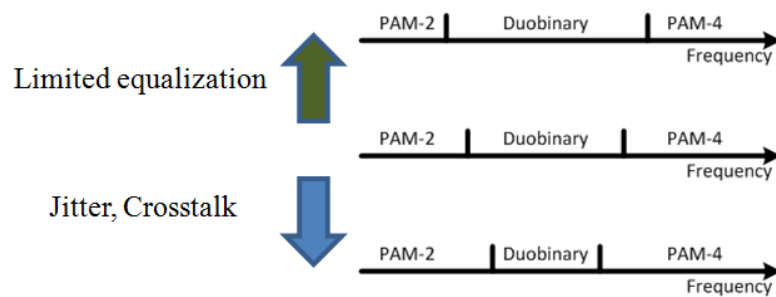


Fig. 23 Modulation selection range changes due to limited equalization, jitter and crosstalk

## CHAPTER IV

### EQUALIZATION TECHNIQUES AND ANALYSIS

High-speed ADC power consumption is the most critical bottleneck in ADC-based receivers to rival binary receivers. ADC resolution requirement as well as its power dissipation is alleviated either by partially equalizing the signal before quantization, or by enhancing digital equalizer capability that offers larger margin to tolerate quantization distortion. This chapter starts with introducing for linear PAE architectures and evaluating PAE performances on ADC resolution requirement reduction by statistical analysis. Then, the advanced power efficient ADC-based receiver architecture is proposed for further saving digital dynamic power consumption. In addition, a potential use of DFE embedded in ADC as a PAE and its drawbacks are discussed.

#### **4.1 Linear Partial Analog Equalizer (PAE)**

Cascaded analog and digital equalization is initially motivated to use of additional DFE following analog FFE due to insufficient equalization with limited analog equalization complexity [34]. As increasing data rate requires more robust equalization in high-speed wired communications, the idea of partial analog equalizer (PAE) is extended for general digital equalization, in which main equalization is performed in digital domain while PAE reduces ADC resolution requirement [12]. Fig. 24 presents signal passed through two cascaded linear equalizers, 2-tap FFE and 6-tap



FFE. Although 2-tap FFE is insufficient to have open eye, dynamic range is impressively limited (Fig. 24(b)) and FSR adjustment allows for relaxing quantization distortion with reduced ADC input dynamic.

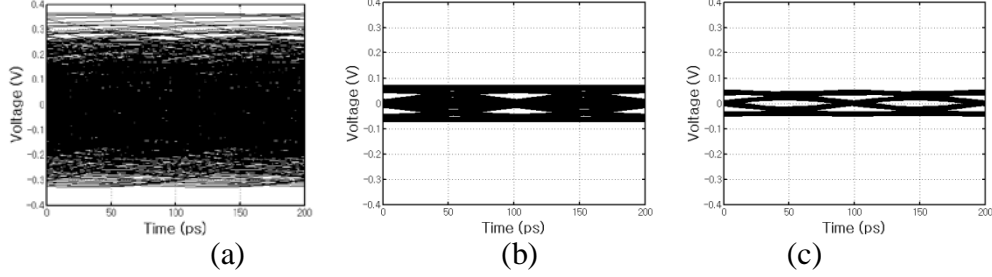


Fig. 24 1k-bits 10Gb/s PAM-2 eye diagram with (a) channel only, (b) channel + 2-tap FFE, and (c) channel + cascaded 2-tap and 6-tap FFE at a channel with -35.1dB loss @5GHz

In this work, 3-tap TX FFE (pre-emphasis) [4] and 2-tap sampled-FFE (SFFE) [35] are considered as linear PAE architectures among different types of existing analog equalization solutions, considering implementation complexity and power overhead. Even though continuous-time linear equalizer (CTLE) is also able to serve as a PAE, CTLE typically has some drawbacks comparing to FFEs. Passive RC CTLE is limited by impedance discontinuity which potentially causes significant reflection. To relief the reflection issue, passive LC CTLE is possibly implemented, but inductive network is not a preferred option for integrated applications [36]. Active CTLE in [20] generates in-band zero and poles controlled by RC source degeneration for high frequency boosting. To obtain sufficient Nyquist frequency gain, transconductance and load impedance

needs to be increased, while large load impedance significantly limits the bandwidth of CTLE itself. As a result, active CTLE requires large biasing current or power consumption. Besides, multiple cascaded stages of CTLE are needed for rapidly attenuating channel, which also increase further power consumption.

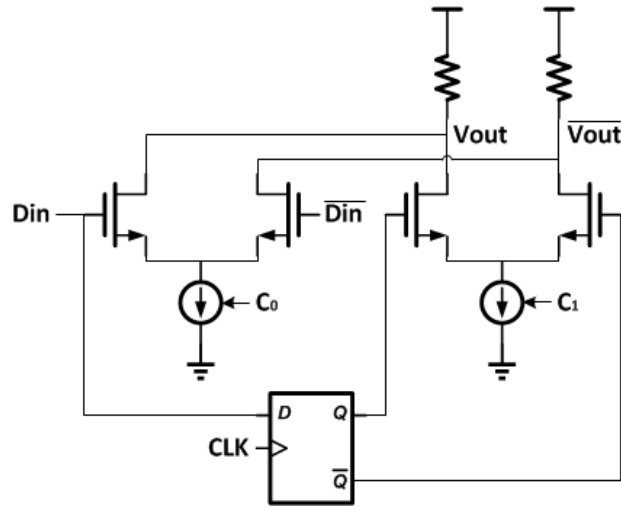


Fig. 25 2-tap TX CML FFE

TX FFE [4] generates controlled transmitting pulses by shared current in multiple TX drivers. Fig. 25 presents an example of 2-tap TX current-mode logic (CML) FFE. Since TX signal peak limits maximum total current of drivers, TX FFE coefficients are normalized by their absolute sum. This means that Nyquist frequency boosting gain to 0dB and corresponding DC loss is unavoidable. Nevertheless, TX drivers sharing total current do not require additional power consumption and there exists a little power overhead for latches and coefficient resolution controls. Another important issue is

limited coefficient resolutions, which restricts typical effective TX FFE complexity to 3-4 taps [2].

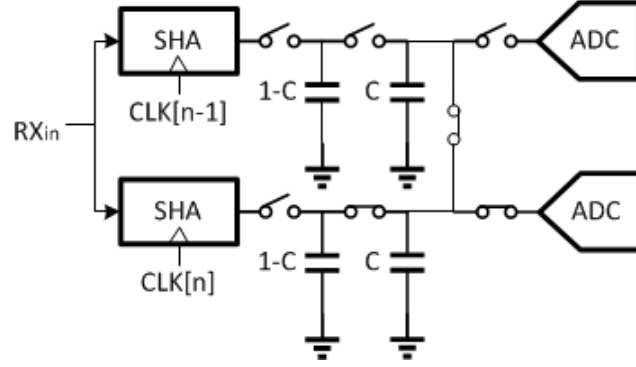


Fig. 26 Simplified 2-tap RX SFFE with time-interleaved ADC

RX-SFFE is also power efficient because it utilizes charge sharing between fractional capacitors used to hold the sampled voltage (Fig. 26) [36]. These charges are sampled by sample-and-hold amplifiers clocked at different phases in time-interleaved ADC. The sampled ADC input voltage with SFFE is

$$v_{ADC,in}[n] = \frac{C}{1+C} v[n-1] + \frac{1}{1+C} v[n]. \quad (4.1)$$

As shown in (19), the coefficients are normalized by absolute sum of two coefficients, which is the same restriction as TX FFE. Since shared charge is not reusable, SFFE is limited to 2-taps. However, it does not require latches and a digital-to-analog converter (DAC) in time-interleaved pipelined ADC or successive-approximation (SAR) ADC is possibly reused for coefficient control, so that SFFE has potentially more power efficiency comparing to TX FFE.

In previous work [7], ADC resolution requirement reduction with PAE is analyzed. This previous work compares required ADC resolutions with sampled FFE and digital DFE to those with digital FFE and digital DFE. However, there is no reason to limit the use of digital FFE with partial analog FFE. One of the most competitive advantages of ADC-based receivers is extensibility of equalizer complexity with large number of FFE taps, which allows for effectively canceling both of pre and post cursor ISI over high channel attenuation. Moreover, the simulated channel losses range from 15.2dB to 34.5dB at 5GHz of Nyquist frequency. Although these results provides a guidance how ADC-based receivers rival to binary receivers in terms of power and area complexity, more interest is higher data rate and channel attenuation, in which ISI control by analog FFE and multi-tap DFE is extremely challenging. This motivates our work to explore PAE performance at 12.5Gb/s data communications over higher loss channels up to 46.4 dB at Nyquist frequency.

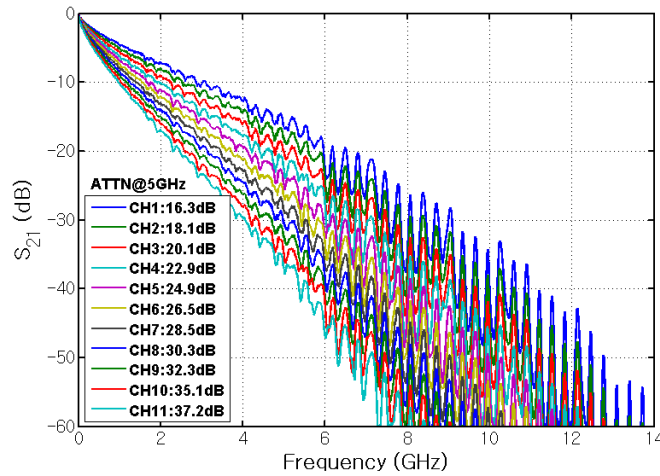


Fig. 27 Channel group 1: channel frequency responses without reflection [18]

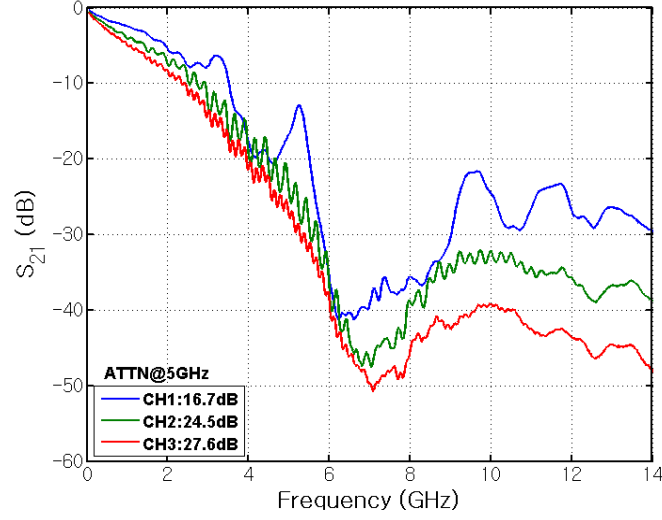


Fig. 28 Channel group 2: channel frequency responses with reflection [18]

Here, the minimum ADC resolution requirements without PAE and with 2-tap RX-SFEE and 3-tap TX-FFE are evaluated by statistical analysis referred to  $1e-12$  of the target BER. Fig. 27 and 28 present two groups of channels used in statistical analysis; group 1 includes smoothly attenuating channels; group 2 contains three different channels with significant reflection. ADC resolution requirement estimation starts with 10Gb/s data rate and then 12.5Gb/s data rate at the same channels with 2-3 taps of FFE PAE and enough digital equalizer complexity is simulated to observe performances under high channel loss. Statistical analysis settings are summarized in Table IX.

Table IX Simulation settings for ADC resolution requirement

Signaling	PAM-2
Data rate	10Gb/s and 12.5Gb/s
Input dynamic	1Vppk
Random noise	1mVrms
RX DJ	2% UI
RX RJ	2% UIrms
Target BER	1E-12
ADC res.	3-8 bits
PAE	No / 2-tap FFE / 3-tap FFE
PAE res.	5 bits
Digital equalizer	4/1-tap FFE/DFE 6/1-tap FFE/DFE 12/2-tap FFE/DFE (reflective channels)
Digital equalizer res.	1 bit greater than ADC res.

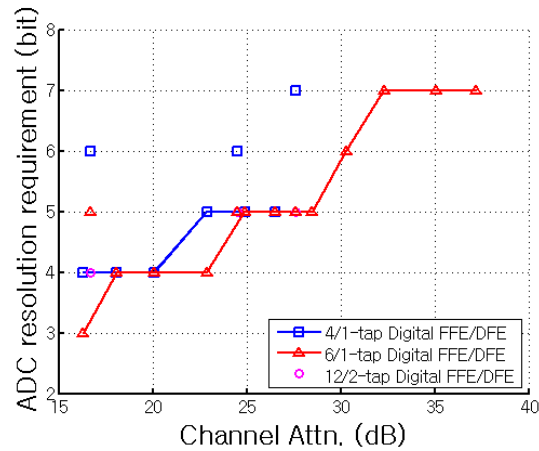


Fig. 29 10Gb/s minimum ADC resolution requirement without PAE

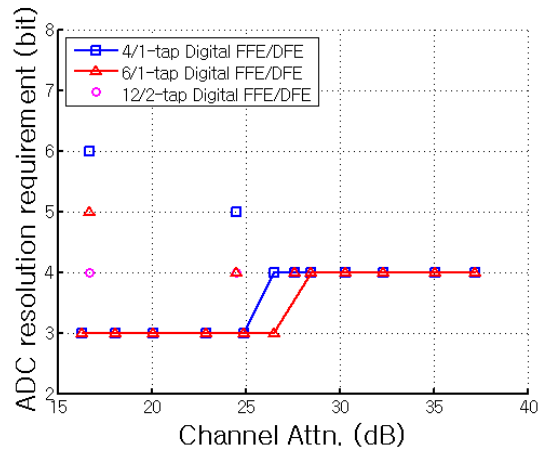


Fig. 30 10Gb/s minimum ADC resolution requirement with 2-tap FFE PAE

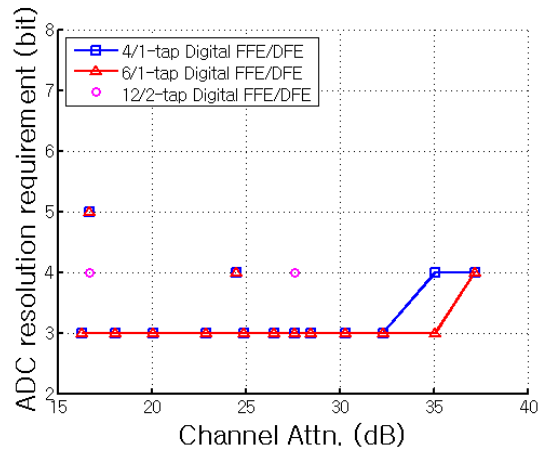


Fig. 31 10Gb/s minimum ADC resolution requirement with 3-tap PAE

Fig. 29 to 31 shows the results of required ADC resolution reduction by PAE at 10Gb/s. Connected data points are from the smoothly attenuating channel group 1 in Fig. 27 and separated points are from reflective channel group 2 in Fig. 28. ADC-based receivers without PAE require high ADC resolutions as shown in Fig. 29, which

discourages to choose ADC-based receivers against binary receivers. Although all the simulated channels can be equalized enough by 3-tap analog FFE and 3-tap DFE, an ADC based receiver with 4-tap digital FFE and 1-tap DFE limitedly tolerates up to 28.5dB of channel loss within 8-bit ADC resolution. To overcome 37.2dB of channel loss, 7-bit ADC and 6/1-tap digital FFE/DFE is required. Besides, channel reflection requires further increasing equalization complexity. 4/1 tap and 6/1 tap digital FFE/DFE are not sufficient for equalizing the three reflection channels and need 1-2 bits more of ADC resolution comparing to gently attenuating channels having similar channel loss.

With PAE, however, ADC resolution requirements are impressively mitigated at smoothly attenuating channels. Without reflection, up to 4 bits of resolution requirement reduction is achieved by 2-tap RX SFFE (Fig. 30) and 3-tap TX FFE accomplishes maximum 5 bits of ADC resolutions saving at over 35dB of channel loss (Fig. 31). An importance observation is that significant reflection causing resonant null at lower frequency than Nyquist frequency (channel 1 in group 2) deadly degrades PAE performance. In this case, increasing digital equalizer complexity improve ADC resolution requirement more efficiently than PAE. Overall, with maximum complexity for both digital equalizer and PAE, only 4-bit ADC resolution is needed for all the channels.



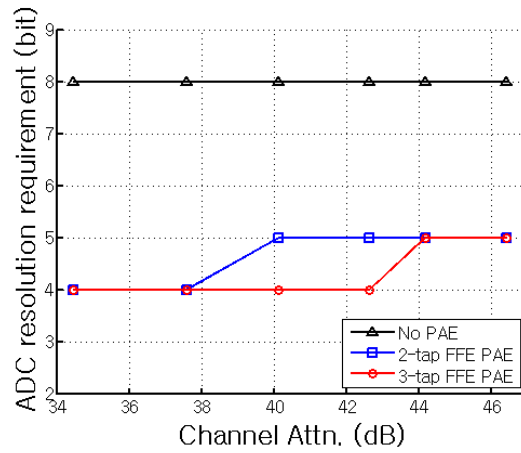


Fig. 32 12.5Gb/s minimum ADC resolution requirement with 6/1-tap digital FFE/DFE and No/2/3-tap PAE at channel 6-11 in group 1

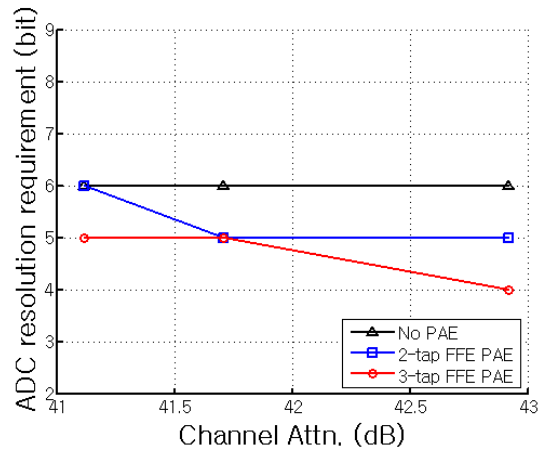


Fig. 33 12.5Gb/s minimum ADC resolution requirement with 12/2-tap digital FFE/DFE and No/2/3-tap PAE at channel 1-3 in group 2

Simulation results with 12.5Gb/s data rate at channel 6-11 in channel group 1 and at channel 1-3 in channel group 2 are presented in Fig. 32 and Fig. 33. In these cases, channel attenuations range from 34.4dB to 46.4dB. Digital equalizer complexities are selected to 6/1-tap FFE/DFE for channel group 1 and 12/2-tap FFE/DFE for channel

group 2 for enough digital equalization to achieve optimal ADC resolution reduction by PAE. The other simulation settings are the same as the previous 10Gb/s cases in Table IX. Note that as channel loss exceeds over 40dB, 12.5Gb/s data suffers from extreme pre-cursor ISI which is uncontrollable by only 3-tap analog FFE. If limited-tap FFE aggressively forces to cancel the pre-cursor ISI, cursor amplitude is significantly reduced and post-cursor ISI gets even worse. ADC-based receiver with 3-tap FFE PAE, meanwhile, requires only 5-bit ADC resolution over all the channels. Considering growing requirement for large number of FFE taps for pre-cursor ISI and reasonable ADC complexity with PAE, it is expected that ADC-based high-speed link will become the main stream in future wired communication with continuously increasing data rate as long as electrical channel is used.

## 4.2 Hybrid Architecture

Although PAE improves ADC power consumption by alleviate resolution requirement, it also limits SNDR due to attenuated cursor amplitude in addition to channel loss and quantization distortion. This leads potential increasing digital equalizer complexity to overcome degraded SNDR. Consequently, power consumption of heavy digital equalizer is going to become as considerable as high-speed ADC power dissipation. Meanwhile at the input of the ADC, a signal does not include quantization distortion even though it is not sufficiently equalized by PAE. This means that signal integrity with PAE at the input of the ADC may not excessively differ from that after fully equalized by digital equalizer.

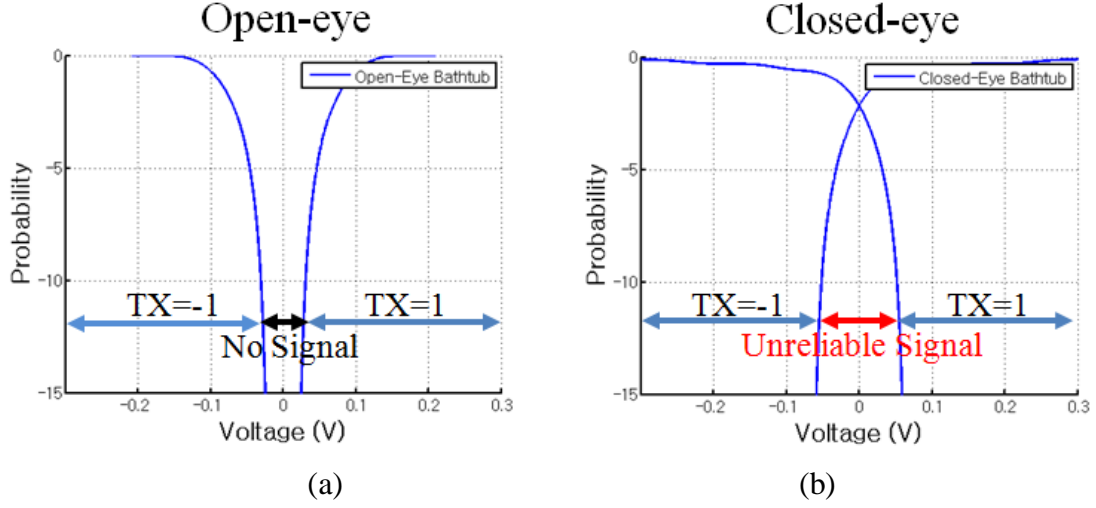


Fig. 34 Statistical bathtub curves (a) with open-eye and (b) with closed-eye

The proposed hybrid architecture is motivated to make the utmost use of the information from the ADC output signal. Fig. 34 presents two statistical bathtub curves with PAM-2 signaling, one has an open eye (Fig. 34(a)) and the other has a closed eye (Fig. 34(b)) referring  $1E-12$  of the target BER. The later may fail to achieve the target BER if decisions are made relying on the zero-threshold; however, there still exists reliable region. For instance, if a received signal is low enough than a certain negative threshold, the probability that is originated from '1' transmitting symbol is lower than the target BER. Likewise, high enough signal most probably comes from '1' symbol without any additional equalization. Therefore, the proposed hybrid receiver defines the region between certain threshold voltages as the unreliable region and selectively equalizes only unreliable signals in digital in order to save dynamic digital power consumption.

There are two possible types of selective equalization, channel-by-channel selection and symbol-by-symbol selection. The former is the case having open eye with PAE in some channels when a transceiver communicates through multiple channels. In this case, a majority of ADC power is saved by limiting ADC to generate only MSB. Although this provides better programmability for multi-channel communications, channel properties are typically known and fixed in wired communications. This more motivates symbol-by-symbol selection.

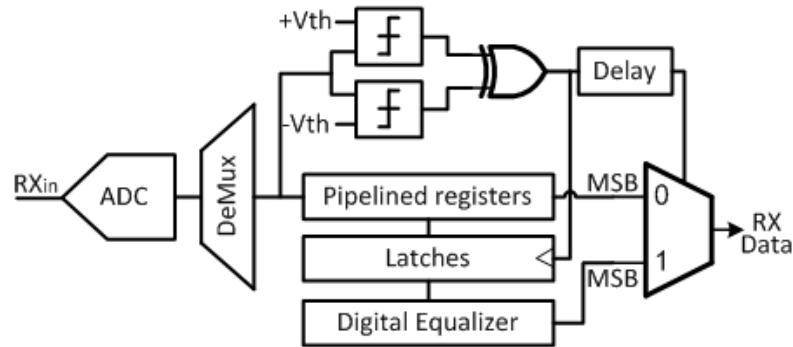


Fig. 35 Proposed hybrid architecture

The proposed hybrid architecture targets high attenuation channels in which PAE does not sufficiently equalizes ISI but enhances ADC resolution requirement. Fig. 35 describes the hybrid receiver. Threshold detector takes the output of ADC and outputs 'high' if the received signal is between thresholds. Then, latches passes received signal in pipelined registers through a digital equalizer and select multiplexer to take the decision from the digital equalizer. Otherwise, digital equalizer input is fixed, so that

there is no dynamic power consumption in digital equalizer. In this case, decision is made by MSB of the ADC outputs in memory. Power overheads are from threshold detector, multiplexer and latches and delay cells to synchronize threshold detector with digital equalizer.

Fig. 36 presents hybrid receiver power efficiency at 10Gb/s with channels in Fig. 27 and 28, in terms of the ratio of disabled digital equalizer versus enabled. Thresholds are obtained by statistical analysis with the same simulation settings in Table IX. Power efficiency is estimated by 1M bits of transient simulation because statistical analysis is not accurate for highly probable events due to correlations between unequalized symbols. Without PAE, power efficiency is rapidly degraded as channel loss increases because of large variance of ISI but PAE dramatically improves power efficiency of hybrid architecture.

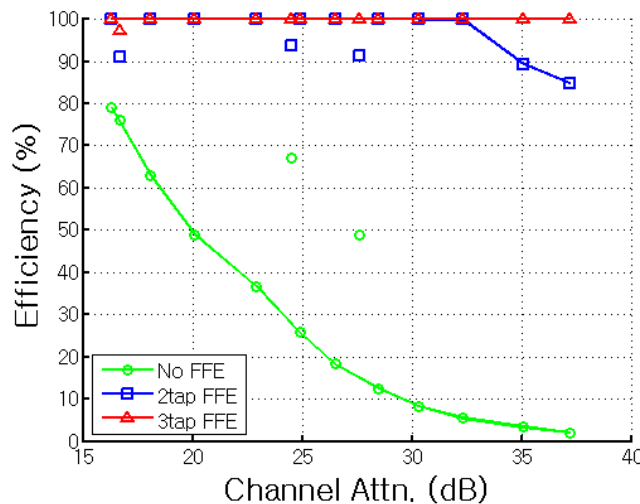


Fig. 36 10Gb/s digital equalization power saving efficiency with No/2/3-tap FFE PAE

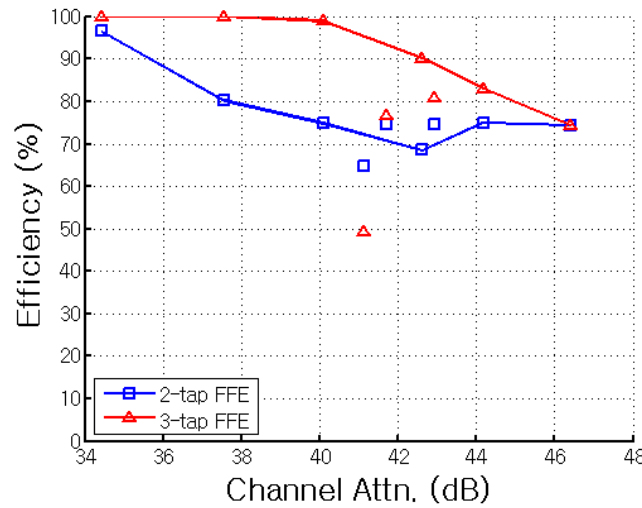


Fig. 37 12.5Gb/s digital equalization power saving efficiency with 2/3-tap FFE PAE

Despite impressive achievement in digital equalization power saving, ADC is always working so that overall power consumption may be much larger than a simple binary receiver with a few taps of additional DFE in 10Gb/s cases. More interestingly, power efficiency in Fig. 37 obtained with 12.5Gb/s data rate still shows impressive power efficiency, even with considerable complexity of required digital equalizer complexity. As mentioned in previous section, in these channels requires 6/1-tap digital FFE/DFE for smoothly attenuating channel and 12/2-tap digital FFE/DFE for channels with significant reflection. In all the ranges up to 46.4dB of channel loss, at least 64% of power efficiency is achieved. Another important finding is that the simulation result at reflective channel 1 implies MMSE does not necessarily results in optimal power saving. It is because PAE FFE coefficients are simply optimized by MMSE algorithm, and insufficient equalization by PAE causes significant residual ISI whose PDF does not

follow a normal distribution. This requires custom optimization to achieve optimal power efficiency, such as parametric sweep by statistical analysis.

#### 4.3 Decision Feedback Equalizer (DFE) for Partial Equalization

A couple of previous works propose embedded DFE into pipelined ADC [37], and SAR ADC [35]. These applications are time-interleaved ADC and utilize MSB of the previously sampled symbol at neighbor time-interleaved ADC as the reference for DFE subtraction. While the embedded-DFE approaches allow DFE to be efficiently implemented with a little power overhead, the use of those ADC output has to be carefully considered. Although the full resolution ADC outputs can potentially be used for FSR or digital CDR, the interest in this work is their possibility to serve as PAE.

Simple calculation of error probability with embedded-DFE and following digital equalizer predicts that these architectures are not efficient for PAE, because DFE subtraction relies on pre-decision at the output of ADC, not the final decision after digital equalization. Assuming that  $p$  is the probability of previous wrong decision at embedded-DFE,  $q$  is final error probability after digital equalization with previous correct decision, and  $r$  is that with previous wrong decision, then the total error probability is,

$$P_e = (1 - p)q + pr. \quad (4.2)$$

Here exists contradiction. Since a wrong decision doubles corresponding post-cursor ISI, both of  $q$  and  $r$  cannot be reduced by fixed digital equalization. Also,  $p$  less than the

target BER means no need of additional equalization. Thus, the overall error probability is significantly limited.

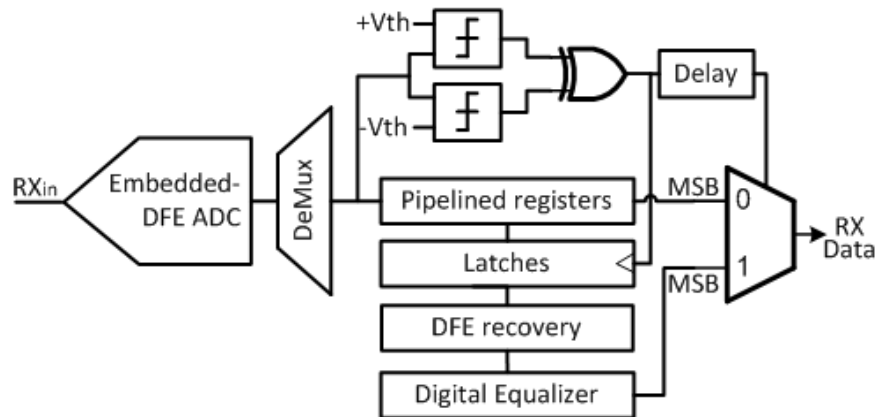


Fig. 38 Modified hybrid architecture for embedded-DFE

There is another possible use of embedded DFE with the modified hybrid architecture in Fig. 38. In this architecture DFE subtraction is recovered based on MSBs of the ADC output symbols in memory and digital equalization is performed. Although the advantage of DFE is not taken for additional digital equalization, reduced unreliable region at the output of ADC allows for increasing chances not to take digital equalization. The key issue for reducing unreliable region is DFE error propagation. If the probability of wrong decision of embedded DFE based on previous correct decision is not sufficient for the target BER but low enough to prevent error propagation, ideal DFE can be assumed and the reduced thresholds can be obtained by statistical analysis. Based on the assumption of ideal DFE without error propagation, power efficiency of hybrid architecture with additional 1-tap embedded DFE in Fig. 37 is recalculated by



statistical analysis. Fig. 39 shows that additional 1-tap DFE achieves maximum 27% of power efficiency improvement comparing to the case without DFE.

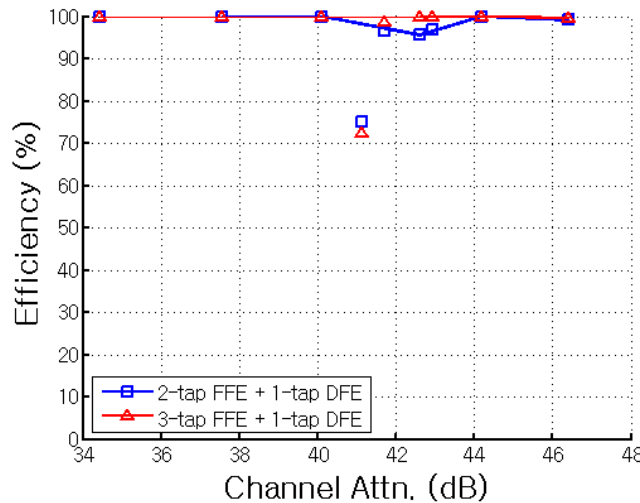


Fig. 39 12.5Gb/s digital equalization power saving efficiency with 2/3-tap FFE and 1-tap ideal DFE PAE

Now, the important two issues are how much pre-decision quality is required for assumption of ideal DFE without error propagation and how the threshold voltage changes if error propagation is not negligible. The most common approach for evaluating error propagation is Markov chain analysis [38]. However, it is not applicable for this case because of exponentially increasing computational complexity with significant amount of residual ISI after PAE. In this work, symbol-based transient simulation with 1Gbits of data is performed and compared to the statistical analysis with ideal DFE assumption to determine minimum requirement of the pre-decision BER at

the output of embedded-DFE ADC for neglecting error propagation. From the transient simulations, no error propagation is observed at all the smooth channels, while critical error propagation degrades power efficiency at all the reflective channels. Fig. 40 presents the transient CDF at the marginal case of propagated error among observations. The result shows that even  $1e-6$  of error rate is not enough for ideal DFE assumption. Again, with the quality signal integrity after PAE, an ADC-based receiver is not competitive versus a binary receiver.

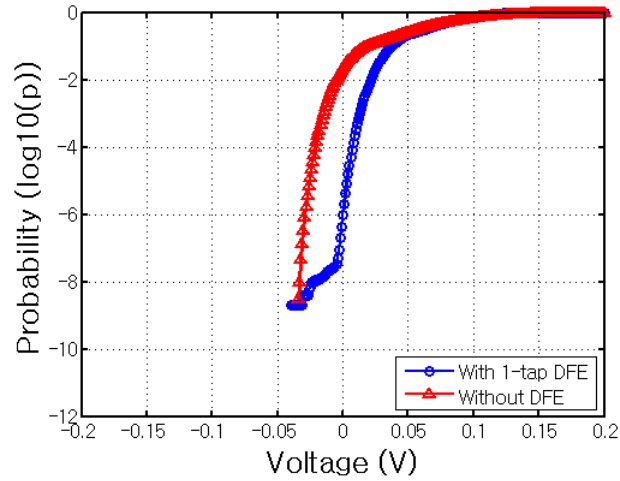


Fig. 40 12.5Gb/s 1G-bit transient CDF at reflective channel 3 with and without 1-tap DFE.

Here is an interesting observation, in which hybrid power saving is enhanced by 1-tap DFE despite error propagation. The simulation is performed with 10Gb/s data rate and 1-tap embedded DFE at the channel 3 in Fig. 27 whose Nyquist frequency loss is 20.1dB. As shown in the transient PDF in Fig. 41, error propagation is

observed and tail probability is extrapolated by polynomial curve fitting to estimate modified threshold with error propagation. Extended tail provides a new threshold to define unreliable region referred to  $1e-12$  of probability. Based on the new threshold with 1-tap DFE, Fig. 41 shows the power efficiency improvement comparing to the case without DFE in Fig. 36. Interestingly, 1-tap DFE reduces the threshold from 0.32V to 0.02V and 27.93% of efficiency improvement is achieved even though error propagation happens. Note that this improvement only shows that the percentage of dynamic power saving by fixing digital equalizer input. Due to additional power overhead for DFE subtraction recovery, the actual power saving will be degraded. The observed power efficiency improvement is because error propagation limits some worst case data patterns for residual ISI. Since a propagated error from the same polarity of previous symbol rather helps the current decision, errors propagate only with limited data sequences.

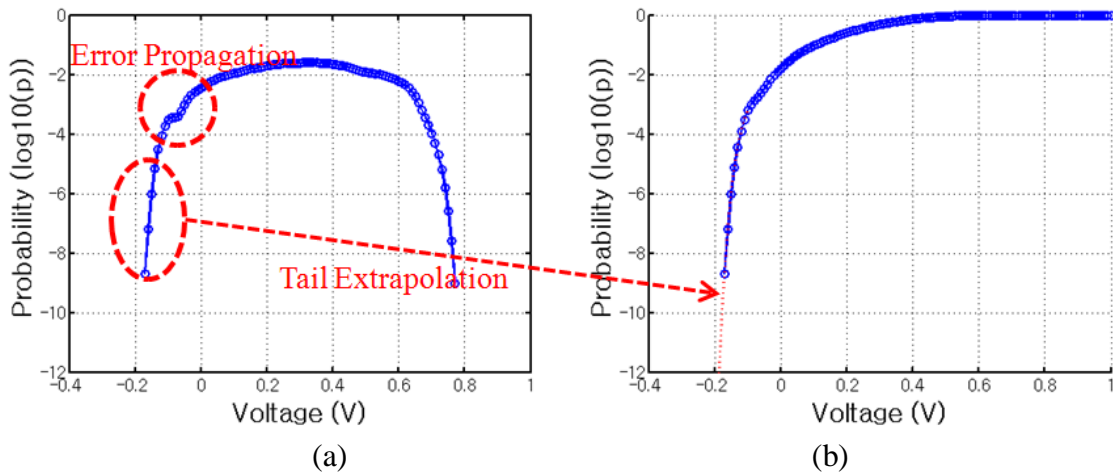


Fig. 41 (a) Transient simulation PDF and (b) its CDF and tail extrapolation with 10Gb/s data rate and 1-tap DFE at channel 3 in Fig. 27.

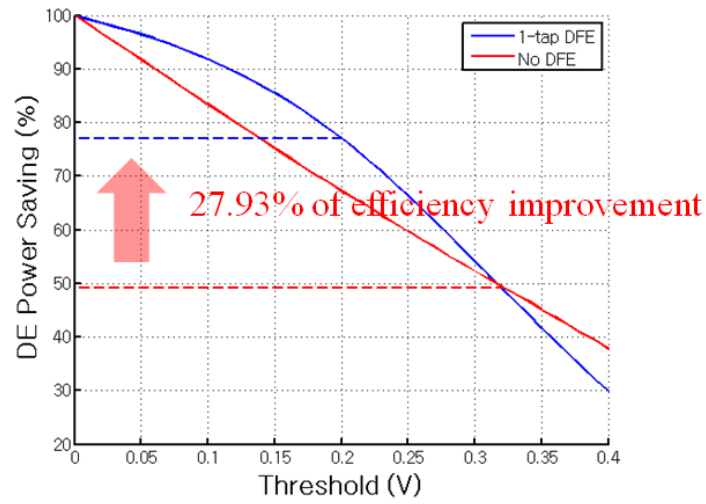


Fig. 42 Power efficiency improvement by 1-tap DFE

In summary, embedded-DFE ADC is not very promising solutions for further digital equalization unless error propagation is well controlled somehow. Although embedded-DFE with error propagation can improve power efficiency in some specific cases, error propagation analysis with insufficiently equalized channels is very challenging and still remains as a future work.

## CHAPTER V

### CONCLUSION AND FUTURE WORK

In conclusion, this work presents a statistical modeling for ADC-based links with digital FFE-DFE equalizers. Three cases of digital equalization with 4-6 bit ADC are simulated and compared with Gaussian approximation based analysis and transient simulation results. The results show that Gaussian approximation is often pessimistic due to unbounded tail probability, but it is also possibly optimistic with large variance of quantization distortion and large number of FFE taps. Proposed modeling better tracks the actual observations in transient simulation and helps avoid potential over-design or system fail to satisfy the low target BER.

ISI control and mitigation can be achieved either by using advanced signaling or by equalization. Since PAM-4 and duobinary worsen eye opening for low-loss channels, modulation scheme should be carefully selected. From the simple profiling of channel attenuation profiling, the best performing signaling can be selected. However, insufficient equalization complexity, jitter and crosstalk possibly change the initial modulation selection. Specifically, limited equalization encourages duobinary selection, while jitter and crosstalk increases the chance to select PAM-4.

Power and area consumption of high-speed ADC is the most critical bottle neck for ADC-based links. However, in future high-speed link with further increasing data rate, extremely increasing demands for equalization more motivates to take the advantages of DSP. PAE is a very promising solution to alleviate ADC resolution

requirement so as to reduce high-speed ADC power and area consumption. This work presents that 12.5Gb/s data communication in channels having up to 46dB of loss requires only 5-bit ADC with 3-tap FFE PAE and 6/1-tap digital FFE/DFE. Even with significant channel reflection, 5-bit ADC with 3-tap FFE PAE and 12/2-tap digital FFE/DFE can endure up to 43dB of channel loss at 12.5Gb/s data rate. Moreover, PAE also improves additional power digital saving with hybrid architecture. With 3-tap PAE, at least 64% of dynamic power saving is achievable in all the cases simulated in this work. Meanwhile, DFE cannot serve as a PAE, but can be used in hybrid architecture with DFE recovery. However, the power saving performance with DFE is significantly limited by error propagation.

Overall, ADC-based link will become a main stream of high-speed transceiver in future wired communication as long as electrical channel is used. Considering ADC-based receivers target high attenuating channel with large data rate and provide robust equalization capability, a combination of PAM-4 and ADC-based receiver is the most promising solution for future high-speed I/O links. In addition, enhanced power efficiency by PAE and hybrid architecture will make this change earlier.

## 5.1 Future Work

This work could be extended to diverse advanced researches. In this section, some of possible extensions of research are presented. First of all, ADC-based link with PAM-4 suffers from additional SNDR degradation due to constellation loss. This may motivate advanced digital equalizer beyond FFE-DFE, such as maximum-likelihood

sequence estimator (MLSE) [19]. However, MLSE is non-linear equalizer so that statistical analysis is challenging with MLSE. Although one possible analysis approach is Markov chain modeling, this typically becomes extremely complicated with large taps of channel modeling with uncontrolled ISI.

Secondly, actual power modeling is not included in this work. It is clear that ADC-based link performs better with high channel attenuation while binary receiver may be preferred for low loss channels. Thus, power break-even point between them in terms of channel loss and data rate needs to be analyzed. Actual power saving from hybrid architecture is still not clear because digital equalizer power itself may not be significant comparing to ADC power, and dynamic power consumption for enabling and disabling digital circuits in hybrid architecture is often critical.

Thirdly, the hybrid architecture with embedded-DFE and error propagation mitigation techniques could be further researched as well as its statistical modeling. Here are examples of previously proposed approaches to relax error propagation. In [39], erasure-DFE is proposed, which is similar idea to hybrid receiver but DFE subtraction is selectively taken based on a certain thresholds. Also, pre-coded data can help to alleviate error propagation [40]. The advantage of coding method is that this does not increase structural complexity of transceivers.

Finally, due to significant channel loss and additional constellation loss with PAM-4, actual implementation of ADC having tiny LSB for optimal FSR could be challenging. Generally, small LSB of ADC is not favorable because of many design issues such as current injection from non-ideal CMOS switch, limited input sensitivity of

clocked comparator and inductive kickback to sampled ADC input. Variable gain amplifiers (VGAs) can help this issue, but limited high frequency gain and bandwidth makes implementation difficult as well as generated noise by VGAs. Thus, resolution requirement reduction is potentially limited by non-ideal FSR adjustment and achievable enhancement in reality could be explored.



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